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July-Dec.

July-Dec., 1928

PROCEEDINGS OF

The Institute of Radio Engineers

Volume 16 July, 1928 Number 7

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GENERAL INFORMATION

The PROCEEDINGS of the Institute is published monthly and contains the papers and the discussions thereon as presented at meetings.

Payment of the annual dues by a member entitles him to one copy of each number of the PROCEEDINGS issued during the period of his membership.

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Walter G. Cady
Manager of the Institute, 1928

Walter G. Cady

MEMBER OF THE BOARD OF DIRECTION OF THE INSTITUTE, 1928

Walter G. Cady was born in Providence, Rhode Island, December 10, 1874. He received the Ph. B. and M. A. degrees from Brown University in 1895–96, and the Ph. D. degree from the University of Berlin in 1900. From 1895 to 1897 Dr. Cady was instructor in mathematics at Brown University. During the period of 1900–02 he was a Magnetic Observer for the Coast and Geodetic Survey. He left this service in 1902 to become an instructor in physics in Wesleyan University at Middletown, Conn. In 1903 he was promoted to Assistant Professor and in 1907 Professor and Head of the Physics Department.

He is a Lieutenant Commander in the U. S. Naval Reserve, a member of the Physical Society, Optical Society, American Institute of Electrical Engineers, and the American Academy of Art and Science.

During the war Dr. Cady was associated with groups at the General Electric Company's Research Laboratory, Columbia University, and the Naval Experimental Station in New London, Connecticut, working on the application of piezo-electricity to the problem of submarine detection.

Dr. Cady has published various papers on cathode rays, magnetic declinograph, electric arc, electric oscillations, piezo-electric resonator and oscillator. In 1923 he made the first direct international comparison of frequency standards by comparing a set of his quartz resonators with frequency standards in Italy, France, England, and the United States.

Dr. Cady has contributed several important papers on the subject of piezo-electricity to the PROCEEDINGS of the I. R. E. Since the organization of the Connecticut Valley Section of the Institute he has been its chairman. Dr. Cady was appointed by the Board as Manager of the Institute for 1928. He is a Fellow in the Institute.

CONTRIBUTORS TO THIS ISSUE

Carson, John R.: Received the B. S. degree from Princeton University 1907; E. E. degree 1909, and M. S. degree 1912. Instructor in physics and electrical engineering, Princeton University, 1912–1914. Engineering Department, American Telephone and Telegraph Company 1914–1917. Engineer, Department of Development and Research, American Telephone and Telegraph Company, 1917 to date. In charge of theoretical transmission studies. Awarded Morris Leibmann Memorial Prize by the I. R. E. in 1924 for contribution to engineering circuit theory and for invention of single side-band carrier suppressed transmission system. Author of some forty technical and scientific papers and a treatise on Electric Circuit Theory and operational calculus. Member of the Institute.

Dellinger, J. H.: (See PROCEEDINGS for May, 1928).

Hanson, Malcolm P.: Born October 19, 1894. Studied mechanical engineering at University of Wisconsin 1914–17 and engaged in radio experimental and military connection work. During the war served in U. S. Navy in charge of several land stations. In 1918 Commander Ensign and attached to Naval Aviation. After the war returned to University of Wisconsin, 1920–24, continuing studies in electrical engineering and radio, and associated with Physics Department in radio experimental work and as Chief Operator of Stations 9XM and WHA. Since 1924 at Naval Research Laboratory engaged in aircraft radio development work. Was in charge of radio arrangements and special design for Commander Byrd's Antarctic Expedition and Transatlantic Flight and has been detailed Chief Radio Engineer to the Byrd Antarctic Expedition. Associate member of the Institute.

Jolliffe, C. B.: Born in West Virginia, November 13, 1894. Received the B. S. degree from West Virginia University 1915; M. S. degree from West Virginia University 1920; Ph. D. degree Cornell University 1922; from 1917 to 1918 and 1919–20 was instructor in Department of Physics, West Virginia University. Instructor in Physics, Cornell University 1920–22. In Radio Section, Bureau of Standards 1922–26 and 1927 to date.

Marrison, W. A.: Born May 21, 1896 at Inverary, Ontario, Canada. Received the B. Sc. degree from Queens University, Canada, and the M. A. degree from Harvard University. During the war was wireless mechanic in Royal Flying Corps. Received the B. Sc. degree in Physics at Queens University in 1920. Received the M. A. degree in mathematics and physics at Harvard University in 1921. Since 1921 has been in the Western Electric Company and Bell Telephone Laboratories working on development of constant frequency sources, studies of quartz crystals as resonators, temperature control, and picture transmission. Member of the Institute.

O'Neill, H. M.: Born September 15, 1899 at Moneton, New Brunswick, Canada. Actively engaged in radio work since 1920. First as an operator and later on test and engineering for the General Electric Company. For the past several years has been engaged in antenna design and investigations of antenna characteristics.

Pratt, Haraden: (See PROCEEDINGS for May, 1928).

Zandonini, Elizabeth M.: Born December 11, 1898 in New York City. Graduate of McKinley Manual Training High School, Washington, D. C. 1918 and National Radio Institute 1918. Instructor of radio in Reconstruction Division of War Department, 1919–20. Owner and operator of amateur radio station 3CDQ, 1921 to date. In radio section, Bureau of Standards, 1921 to date.

INSTITUTE ACTIVITIES

PROFESSOR ZENNECK COMING TO THE UNITED STATES

Word has just been received by the Institute that Professor Jonathan Zenneck of the Technical High School, Munich, Germany, will attend the September 5th meeting of the Institute in New York City to receive the Institute Medal of Honor and to deliver a paper.

Special arrangements are being made for this meeting. Complete details will be announced in the next issue of the PROCEEDINGS.

JUNE MEETING OF THE BOARD OF DIRECTION

At the June meeting of the Board of Direction, held in the office of the Institute on June 6th, the following were present: Arthur Batcheller, Ralph Bown, W. G. Cady, J. H. Dellinger, A. H. Grebe, R. A. Heising, Donald McNicol, and J. M. Clayton.

The following were transferred or elected to the higher grades of membership: Transferred to the grade of Fellow, Stuart Ballantine and Lewis M. Hull. Transferred to the grade of Member: I. J. Adams, Edward L. Bowles, J. E. Brown, Glenn Browning, John C. Burchard, C. W. DeRemer, John DiBlasi, Frank B. Falknor, James L. Finch, Harry W. Houck, James C. McNary, Earle M. Terry, and Harold A. Wheeler. Elected to the grade of Member: A. M. Cheftel, W. C. Evans, W. A. Marrison, J. C. Randall.

One hundred and sixteen Associate members and twenty Junior members were elected.

J. H. Dellinger was appointed to the Advisory Committee of the United Engineering Society to co-operate in the policy of this organization in enlarging the scope of its Engineering Index.

COMMITTEE ON INSTITUTE AWARDS

A standing committee to consider candidates for the two annual Institute awards, the Morris Leibmann Memorial Prize and the Institute Medal of Honor, was authorized by the Board of Direction at its last meeting. A committee composed of the following members of the Institute has been appointed: J. V. L. Hogan, (Chairman), L. W. Austin, Ralph Bown, W. G. Cady, and A. Hoyt Taylor.

This committee will make recommendations to the Board of Direction for the 1928 Morris Leibmann Memorial Prize.

Members of the Institute having any candidates in mind should communicate with any member of this committee before September 1st.

PRELIMINARY 1928 STANDARDIZATION REPORTS

The preliminary draft of report of the Institute's Committee on Standardization for 1928, as submitted by the seven subcommittees, is now available in printed form.

This preliminary report has been prepared for submission to the main Committee on Standardization for final action. It is a document of some one hundred pages.

Members of the Institute who are interested in standardization should receive a copy of this preliminary draft. Copies will be mailed free of charge upon application to the Institute office; however, only a limited supply is available. The comments and criticisms of all members are cordially solicited.

FOREIGN HIGH-FREQUENCY ASSIGNMENT

The attention of the Institute membership is called to the Radio Service Bulletin, No. 133, dated April 30th, 1928, in which will be found a compilation of high-frequency assignments in foreign countries in which is shown the location of stations, call letters, and remarks as to type of service. A copy of this bulletin may be obtained from the Superintendent of Documents, Government Printing Office, Washington, D. C. for five cents.

CHANGE OF POLICY IN HANDLING APPLICATIONS FOR MEMBERSHIP

At the meeting of the Board of Direction held on June 6th, the following change in the policy of handling applications for transfer or election to *any* grade of membership in the Institute was adopted. The Committee on Admissions will examine, monthly, all applications for transfer or election to any grade of membership. The names of those applicants whose applications have been approved will then be published in the next issue of the PROCEEDINGS, which will be in the mail in sufficient time to allow all members of the Institute to advise the office of the Secretary if there are any objections to the transfer or election of any of the applicants listed.

At the meeting of the Board of Direction following the publication of the PROCEEDINGS, these applicants will be considered by the Board of Direction and appropriate action taken.

This new routine will materially increase the time required by the Board of Direction acting upon an application, the normal time, it is expected, being approximately two months from the date of receipt of application by the Institute.

The August issue of the PROCEEDINGS will contain the first list of such applications.

Institute Meetings

NEW YORK MEETING

At the meeting of the Institute held in New York City on June 6th, in the Auditorium of the Engineering Societies Building, 33 West 39th Street, the following papers were presented: "Aircraft Radio Equipment," by Malcolm P. Hanson of the U. S. Naval Research Laboratory; "Radio Aids to Air Navigation," by J. H. Dellinger and Haraden Pratt, presented by Mr. Pratt; a two-reel motion picture illustrating some of the developments carried on by the Army in directive radio transmission and reception, shown through the courtesy of Captain William H. Murphy of the Signal Corps Radio Laboratories, Ft. Monmouth, New Jersey.

The two papers are printed elsewhere in this issue. A limited supply of copies in pamphlet form are available to members of the Institute and can be obtained upon application to the Secretary.

Following the presentation of the papers and the motion picture film, the following took part in the discussions: Ralph Bown, Stuart Ballantine, William H. Murphy, J. H. Dellinger, M. P. Hanson, and others.

Over five hundred members of the Institute and guests attended this meeting.

ATLANTA SECTION

A meeting of the Atlanta Section was held on May 9th in the Ansley Hotel, Atlanta, Ga. D. C. Alexander, vice-chairman of the Section, presided. F. H. Schnell, of the Burgess Battery Company at Madison, Wisconsin, presented a paper "Experimental High-Frequency Radio for Aircraft." In the paper two transmitters were described, one for ground and the other for aircraft. A very compact aircraft receiver for high-frequency work was also described in detail. The speaker described the many experiments conducted by the Burgess Battery Company in the development of this equipment. One model of the aircraft transmitter and receiver is said to weigh only fifty-three pounds including the power supply.

C. F. Daugherty, engineer in charge of Station WSB, arranged to broadcast this meeting through WSB. The talk by Mr. Schnell was also broadcast over portable Station 4 PAG on 85 meters, special permission having been obtained from the Federal Radio Commission.

Seventy-five members and guests attended this meeting of the Section.

Following the paper, the announcement of the election of 1928 officers of the Atlanta Section was made as follows: Walter Van Nostrand, chairman; D. C. Alexander, vice-chairman; and George Llewellyn, secretary.

The next meeting of the Atlanta Section will be held on September 13th at which time Irving Wolff of the Radio Corporation of America will deliver a paper "Sound Measurements and Loudspeaker Characteristics."

BUFFALO-NIAGARA SECTION

On May 9th a meeting of the Buffalo-Niagara Section was held in Foster Hall, University of Buffalo. L. C. F. Horle presided. Dr. L. G. Hector, of the University of Buffalo, presented a paper "Measuring Instruments." Professor Hector outlined a philosophy of measuring and applied this to the case of electrical measuring instruments. All different types of movements and their uses were taken up by Professor Hector, and manufactured examples were exhibited. Professor Hector illustrated with laboratory apparatus the various principles involved in the different types of movements. He also pointed out the differences and likenesses between various types of meters.

Thirty-five members were present at this meeting.

On June 13th a meeting of the Buffalo-Niagara Section was held in Reickerts' Tea Room, 190 Delaware Avenue, Buffalo. L. C. F. Horle and I. R. Lounsberry presided.

This meeting was the annual meeting of the Section and was preceded by a dinner. W. K. Powell presented a talk on "Traffic."

The election of officers of the Buffalo-Niagara Section for the coming season took place. L. C. F. Horle was elected chairman; L. G. Hector, vice-chairman; and C. J. Porter, secretary-treasurer. The reports of the various committees of the Section were read and approved by the members present.

Thirty-eight members of the Section attended this meeting.

CONNECTICUT VALLEY SECTION

In Scott Laboratory, Middletown, Connecticut, a meeting of the Connecticut Valley Section was held on June 1st. K. S. Van Dyke, vice-chairman of the Section, presided. He presented a paper, "Some Recent Experiments with Quartz Spheres," in which he described a number of very interesting phenomena which take place when quartz spheres are made to resound at their natural or harmonic frequencies. The experiments were illustrated with lantern slides and motion pictures. Following the talk, the meeting adjourned to one of the rooms of the laboratory in which Dr. Van Dyke gave demonstrations.

LOS ANGELES SECTION

On May 21st a meeting of the Los Angeles Section was held in the Elite Cafe, 633 S. Flower Street, Los Angeles. D. C. Wallace, chairman of the Section, presided. Two papers were presented. The first by J. J. Jackosky on "Geophysical Prospecting." The second paper, by A. A. Barber on "Electrical Measurements", was illustrated by means of a motion picture film loaned by the Weston Electrical Instrument Corp. Seventy members of the Section attended this meeting.

The next meeting of the Section will be held in the Elite Cafe, September 17th.

PHILADELPHIA SECTION

A meeting of the Philadelphia Section was held in the Franklin Institute on June 1st. J. C. Van Horn, chairman of the Section, presided. Edgar H. Felix presented an illustrated paper on the "Rayfoto System of Picture Broadcasting."

A general discussion followed the presentation of the paper. One hundred and forty members and guests attended this meeting. The next meeting of the Section will be held on June 22nd in the Franklin Institute.

PITTSBURGH SECTION

In the Assembly Room of the Fort Pitt Hotel, Pittsburgh, a meeting of the Pittsburgh Section was held on May 11th.

W. K. Thomas, chairman, presided. Dr. Phillips Thomas, of the Westinghouse Electric and Manufacturing Company, presented a paper, "Two Meter Tube." The paper described a two-meter oscillator connected in the Hartley circuit. Location of the nodes was illustrated through the use of neon-argon indicators, and their resonant indicators employing flashlight bulbs. The possibilities of the construction of two-meter tubes for the generation of 50 to 100 kilowatts of power were discussed. A speculative analysis of the generation of 50 to 100 kilowatts of power at 10 to 20 centimeters and the use of reflectors for such wavelengths were given.

Thirty-seven members attended this meeting.

The next meeting of the Pittsburgh Section will be held on September 18th in the Fort Pitt Hotel.

SEATTLE SECTION

On April 28th a meeting of the Seattle Section was held in the Telephone Building of Seattle. W. A. Kleist, chairman of the Section, presided. A paper by W. V. Sylvester and H. E. Renfro on "Communication Systems of the City Light Department" was presented. Mr. Sylvester gave a talk on the requirements of a communication system for use in conjunction with a power supply plant, pointing out that reliability was highly important in connection with load despatching and other business. The necessity for communication with generating plants located in isolated places was also mentioned. Problems in the design and maintenance of such a system were covered, with a summary of the troubles encountered, such as noise from induced voltages to ground, hazards to life in time of power-line failure, etc. Methods of protection both to the plant and the personnel were described in detail.

Mr. Renfro described the radio telephone system used between Seattle and the hydro-electric plant on the Skagit River, one hundred and twenty miles distant. He also discussed the possibility that transmission was taking place partly over the power line and partly through space since the antennas at both ends are located near the power line. The problem of signalling over this system, it was pointed out, has to be further improved before it can be called entirely satisfactory.

WASHINGTON SECTION

A meeting of the Washington Section was held in Picardi's Cafe, 1417 New York Avenue, N. W., Washington, D. C. on May 10th. Forty-three members attended the dinner preceding the meeting. F. P. Guthrie, chairman of the Section, presided.

A paper by Malcolm P. Hanson on "Aircraft Radio Installations" was presented by Mr. Hanson, and another paper by J. H. Dellinger and Haraden Pratt, of the Bureau of Standards, on "Radio Aids to Air Navigation" followed. In the discussion which followed, C. P. Mirick, A. H. Taylor, and others took part.

Sixty-eight members and guests attended the meeting.

Both papers are printed elsewhere in this issue.

Committee Work

1929 CONVENTION COMMITTEE

A Preliminary Committee on the 1929 Convention was appointed by the Board of Direction and the Washington Section of the Institute with membership as follows: J. H. Dellinger, Chairman, Ralph Bown, J. M. Clayton, A. Hoyt Taylor, Guy Hill, S. S. Kirby, F. P. Guthrie, C. B. Jolliffe, and Alfred Crossley.

A meeting of this Preliminary Committee was held in Washington on May 22nd. Tentative plans for the 1929 Convention were drawn up with appropriate recommendations to the Board of Direction.

The Board of Direction of the Institute approved the report of this Preliminary Committee and authorized the appointment of the following chairmen of the various committees: Executive Committee, F. P. Guthrie; Registration and Arrangements, Alfred Crossley; Fellowship, Professor E. C. Kennely; Trips, S. S. Kirby; Banquet, F. P. Guthrie; Publicity, R. D. Heinl.

A Preliminary Committee to consider entertainment of ladies present was appointed as follows: Miss E. M. Zandonini, Miss Mary T. Loomis, Mrs. F. P. Guthrie, and Mrs. J. H. Dellinger.

COMMITTEE ON ADMISSIONS

A meeting of the Committee on Admissions was held at 1:30 P.M. on June 6th in the office of the Institute. The following members were present: R. A. Heising, Chairman; H. F. Dart, and E. R. Shute. The Committee considered cases of thirty-five applicants for transfer or election to the higher grades of membership in the Institute.

COMMITTEE ON MEMBERSHIP

A meeting of the Committee on Membership was held in the Fraternities Club Grill, 38th Street and Madison Avenue on July 6th. The following members of the Committee were present: H. F. Dart, Chairman, I. S. Coggeshall, and H. B. Coxhead. A number of problems involved in the increase of membership in the Institute were discussed.

COMMITTEE ON SECTIONS

A meeting of the Committee on Sections was held on May 18th in the Institute Headquarters. The following were present: Donald McNicol, Chairman; E. I. Green, M. Berger, and E. I. Shute. The Committee approved the petition of members residing within the territory of New Orleans, La. for the chartering of a New Orleans Section.

Correspondence looking to the formation of Sections in St. Louis and Cincinnati is being undertaken by the Committee.

Mr. Shute gave a report of his trip to the Pacific Coast where he visited the Seattle and San Francisco Section officers.

CHARACTERISTICS OF CERTAIN BROADCASTING ANTENNAS AT THE SOUTH SCHENECTADY DEVELOPMENT STATION*

BY

H. M. O'Neill

(Radio Engineering Department, General Electric Company, Schenectady, New York.)

Summary—*In this paper, the characteristics of broadcast antennas of various standard forms are discussed and experimental data presented on actual installations at the South Schenectady Developmental Station. The effect on signal strength, as measured locally, of varying the antenna height is considered; and also the effect of high steel towers on antennas operated at 380 meters wavelength. Antenna ground system losses, local radiation losses, antenna directional effects and field distortion are also discussed to a limited extent.*

THE presentation of a few experimental results with the modifications of theory suggested thereby may be helpful in obtaining a clearer view of the complex action of broadcast antennas. The following paper is a summary of such results and is based on several investigations conducted principally at the South Schenectady Developmental Radio Station of the General Electric Company. It frequently happened during these investigations that what was meant for the treatment of general problems turned out to be the examination of special cases. This, however, is due to the extreme sensitiveness of radiating systems to external influences; and, as such influences are active to some extent in the average installation, it may be that the most practicable way of arriving at a satisfactory basis for antenna design is through the accumulation of data such as are presented in this paper.

Special terms are used to denote the quantities having to do with antenna characteristics. Principal among these terms are: radiation efficiency, total effective resistance, antenna effective current, radiation resistance, antenna effective height, meter amperes, and radio field intensity. All of these terms are defined in the I. R. E. Report of the Committee on Standardization for 1926, to which reference should be made if necessary for clearer understanding of what follows.

Radiation efficiency is generally accepted as being the prime requisite of a broadcast antenna; but, in a wider sense, the most desirable broadcast radiator is that which most consistently

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* Original Manuscript Received by the Institute, May 8, 1928.

serves the greatest number of receiving stations. We therefore desire not only that the antenna be an efficient radiator but that the energy radiated be directed so that it ultimately becomes available at distant receivers, and also that it traverse the intervening distance with the least possible attenuation. For this reason, while much of the following discussion concerns radiation efficiency, propagation problems (with the exception of fading) are also considered, since they are definitely associated with the problem of obtaining a truly effective broadcast antenna.

Radiation efficiency is defined as: "The ratio of the power radiated to the total power delivered to the antenna at a given frequency." Total effective resistance and antenna effective current are directly measurable by standard methods and completely determine the total power delivered; but what means have we of measuring the power radiated? From the definitions of the quantities, radiation efficiency may be expressed by the ratio of radiation resistance to total effective resistance; but, unfortunately, radiation resistance is hardly a measurable quantity because it cannot be varied to any extent in a broadcast antenna without at the same time varying the loss resistance in an unknown manner. Three other possible avenues of approach to the problem of radiation efficiency still remain; radio field intensity, antenna effective height, and meter amperes.

The intensity of the radiation field at a distant point can, in general, be increased only by increasing either the amount or concentration of the power radiated; and it is not a function of antenna height whether "effective" or actual, except in so far as the antenna height governs the ratio of radiation to total resistance or alters the distribution of the field. Radiation resistance is no more difficult to estimate or to calculate than effective height; and since it is much more useful in predicting results on a power basis—the logical one—it will be used in the following discussion to the exclusion of roundabout reasoning involving effective height. Meter amperes also are of little value in arriving at broadcast antenna efficiency, because for the determination we must calculate or guess the two components of total effective resistance; and this done, with the efficiency established, what need have we of meter amperes?

It can be easily demonstrated experimentally that field intensity measured at a distance of several wavelengths from the antenna varies directly as the antenna current and consequently as the square root of the power, both total and radiated. Therefore, if the average is taken of a number of squared values of field intensity measured on a circle of suitable radius with the antenna as center, then this average should be directly proportional to the power radiated through a zone close to the earth's surface. By measurements of this kind made at a constant radius and referred to a common power basis, we can compare the radiation efficiencies of various broadcast antennas. This is subject to the limitation that we cannot be certain, in comparing antennas of different forms, to what extent an increased average intensity, for a given power input, is due to a greater concentration of power close to the horizon rather than to increased radiation efficiency. The measure of radiated power obtained in this manner will, in the following discussion, be expressed in arbitrary units, since in the author's opinion the present radiation formulas and field-intensity standard are not sufficiently exact to justify our attempting to express the result in watts. A description of the instrument used and details of the method of local field intensity measurement are to be found in a paper entitled "Main Considerations in Antenna Design," by N. Lindenblad and W. W. Brown.[1]

The method of radiation measurement outlined in the previous paragraph requires some standard for efficiency comparison. The regular WGY antenna, erected over one of the buildings in the Schenectady Works of the General Electric Company, was adopted as such a standard for the antennas constructed at South Schenectady. This antenna consists of a horizontal flat top 200 ft. long with a tuned downlead at each end, suspended from two 165-ft. self-supporting steel towers, with a height of flat top above the counterpoise of approximately 120 ft. This antenna is of the multiple-tuned type, but due to the comparatively high ratio of antenna height to operating wavelength the full advantage of multiple tuning is not realized in this particular installation. The antenna dimensions and fundamental wavelength (359 meters), indicate that this structure consists essentially of two inverted "L" antennas placed end to end. Since the ground resistance of each antenna is due to the expansion and contraction of its own induced field, it does not appear that the two existing ground paths are in multiple to any greater extent than it is possible for the electric field of one antenna element in collapsing

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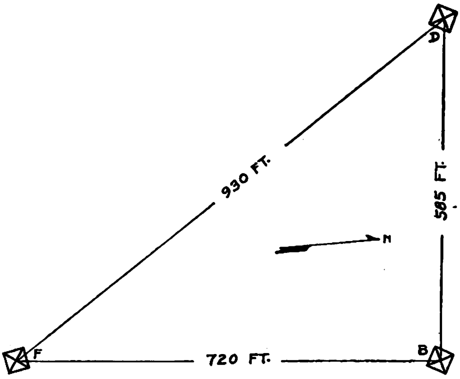
[1] Presented before New York meeting of the Institute, March 3, 1926. PROC. I. R. E., 14, 3, p. 291; June, 1926.

tached to tower *D* and was located over the main building, about 100 ft. to the south of the tower. The efficiencies of both these antennas were remarkably low as compared to WGY, and in addition to this, the field intensity patterns indicated an unusual amount of distortion. From the comparison given by Fig. 3, it will be observed that the field patterns have the same general form and that the distortion in each case bears the same relation to the direction of the line joining the antenna and tower *D*, thus indicating that the tower was resonant at or near the operating frequency and was, therefore, responsible for this distortion and possibly also for the low radiation efficiency.



ARRANGEMENT OF 300 FT. TOWERS
SOUTH SCHENECTADY RADIO DEVELOPMENTAL STATION

Fig. 2.

The fundamental wavelength of the vertical antenna was 334 meters, and the resistance at 380 meters 11 ohms. The radiation was 31 units which was extremely low when compared with the 290 units of WGY. As tower *D* was apparently causing trouble a new vertical of the same dimensions as the first was installed with an improved wire ground in the center of span *D-F*, in which position it was nearly equi-distant from all three of the high towers. This antenna had a fundamental wavelength of 360 meters, a resistance at 380 meters of 34 ohms and a radiation at 5 km of 180 units: which was a decided improvement over the first vertical but still far below the 290 units of WGY.

The height of the vertical in span *D-F* was raised to 235 ft. in accordance with the assumption that higher efficiency would accompany increased radiation resistance. This change resulted in

Two practicable means whereby resonant steel towers may be detuned are: the use of insulators in the base of each tower leg; and the simple expedient of attaching wires at the tower top to produce the equivalent of an umbrella antenna. Base insulators lower the resonant wavelength in a manner similar to the action of a series condenser in an antenna circuit; while on the other hand, the effect of the umbrella of wires attached to the tower top is to increase the resonant wavelength. The value of either device will depend on the previously existing relation between tower and antenna wavelengths. Both of these methods were used in trying to increase the efficiency of the "T" antenna in span *D-F*. The base of tower *D* was insulated with special porcelain insulators; and the effects of towers *B* and *F* were limited by detuning wires applied to their tops. By these adjustments to the towers, the antenna resistance at 380 meters was reduced to 63 ohms and the radiation increased from 210 to 280 units.

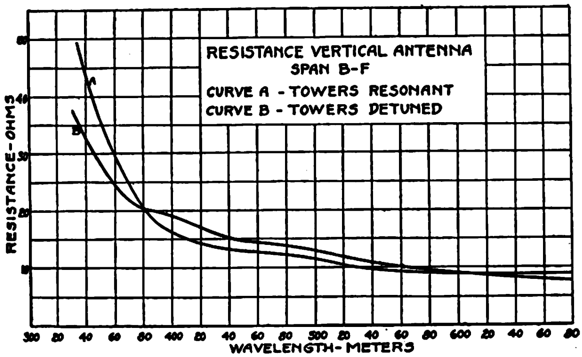


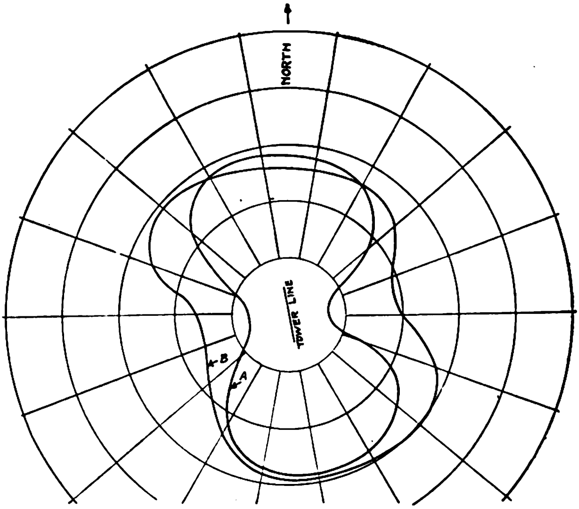
Fig. 4.

Until this time the best that could be done with a 235-ft. antenna using 300-ft. towers was to make its efficiency nearly equal to that of a 120-ft. antenna suspended between 165-ft. towers. It subsequently became necessary to erect in span *B-D* an antenna for high power broadcasting. The antenna erected was a vertical 240 ft. long, very similar to that used in span *D-F*, equipped with a 240-ft. diameter radial wire counterpoise, and having a fundamental wavelength of 374 meters. While the resistance of this new vertical at 380 meters was only 23 ohms as compared with 41 ohms for the vertical in span *D-F*, the field intensity per ampere of antenna current was almost exactly the same in each case. This indicates that the radiation resistance was the same for both, and that the reduced resistance in span *B-D* was entirely due to decreased loss. In consequence of this reduced loss the radiation was increased to 370 units.

The above results were obtained, of course, with the tower wavelengths adjusted; but even with this improvement in antenna efficiency, tower effect still existed, as was apparent by a slight elongation of the field pattern along the tower line. The effect of returning the towers to their original condition of resonance was to distort the field pattern into a pronounced figure 8 along the tower line, and to cut the radiation in half. Nevertheless, the measured resistance at 380 meters was still 23 ohms, although it fluctuated widely while the tower wavelengths were being changed. Apparently, the changes in radiation and loss components compensated for one another so as to leave the total resistance unaffected at this specific wavelength; and at the same time reduced the ratio of radiation to total resistance to one half its former value.

The peculiar condition just considered was to some extent paralleled in a simple vertical antenna erected at the center of span *B-F*. Antenna resistance curves made with and without detuning wires on the supporting towers are shown by Fig. 4, from which it will be seen that, while generally different, the curves cross at 380 meters. Field patterns for the two conditions of tower adjustment are shown by Fig. 5: on which curve *A* is the pattern with the towers resonant and curve *B* the pattern with towers detuned. The difference in apparent efficiency should be directly proportional to the difference in pattern area because both patterns were made with the same antenna power, current, and adjustments. Only two 300-ft. wires attached to the top of each tower and extending downward at an angle of 45 deg. were used in producing this pattern difference, but additional wires brought about no further improvement.

The most important loss in broadcast antenna unaffected by neighboring metal structures takes place in the ground path of the currents caused by the rising and falling electric field. The time-distance idea involved in the explanation of electro-magnetic radiation suggests that an increase in the radiation field for a given antenna current is a consequence of the extension of the field to a greater average distance from the antenna, therefore probably—though not necessarily—increasing the average length of ground path and consequently increasing the ground resistance. Since efficiency depends on the ratio of radiation to total resistance, if the radiation resistance is numerically large, a small numerical increase in loss resistance with increasing antenna height would suffice to hold the efficiency constant. Considerations of this nature led to an investigation at the Developmental Station of the variation in radiation efficiency with antenna height.



FIELD PATTERNS OF VERTICAL ANTENNA IN SPAN B-F
PATTERN A- WITH TOWERS RESONANT
PATTERN B- WITH TOWERS DETUNED

Fig. 5.

The antennas investigated were simple vertical and "T" cage antennas of heights varying from 80 to 240 ft. erected in span *B-F* with the towers detuned. The horizontal cage sections were 2 ft. in diameter and 80, 160 and 240 ft. in length. All antennas used the same counterpoise which was 240 ft. in diameter, consisting of 28 radial wires. Radiation measurements were made on all antennas by the method described early in this paper. The radius of measurements was one mile, the wavelength 380 meters, and the antenna power input 200 watts for all measurements. The arbitrary basis for field intensity values was not the same as for previous measurements; hence there can be no direct comparisons of efficiencies between this group of antennas and those previously measured.

The results of the measurements on the simple vertical and the 160 ft. "T" antenna are shown by Fig. 6. It will be seen that the curve for the 160-ft. cage is practically a horizontal line, while that for the vertical indicates a rapid increase in efficiency with height until a height of 200 ft. is reached. Beyond this point, the curve flattens out although the antenna fundamental was still below 380 meters. The range of fundamental wavelengths for the simple vertical was 135 to 310 meters and that for the 160-ft. flat top 320 to 535. The two curves shown might therefore be plotted in a single continuous curve of efficiency against fundamental wavelength; and this latter curve would be similar to what we might predict would be obtained by progressive increases in the height of the vertical up to 410 ft. The curves for the 80 and 240-ft. horizontal lengths, though somewhat more irregular than for the 160-ft. length, indicated little variation in efficiency with height but showed a slight maximum in the vicinity of 150 ft.

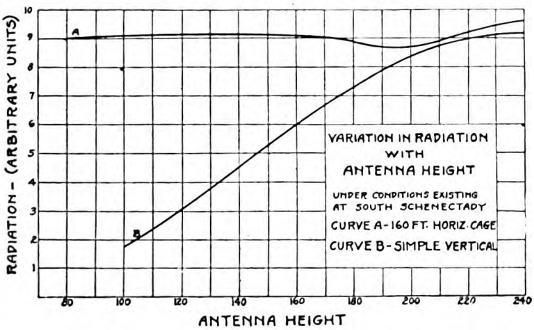


Fig. 6.

It is very important to bear in mind that curves such as those of Fig. 6 must not be taken as a general indication of what may be expected in the average case. Unless measurements of this kind are checked by similar data taken under different conditions of soil resistance, tower effect, etc., general conclusions drawn therefrom are unjustified.

An 80-ft. antenna with a 160-ft. horizontal length, under the conditions existing at South Schenectady, had shown as high a radiation efficiency locally as a 240-ft. vertical antenna; but would this condition also hold great distances? As a preliminary step in determining this point distant comparisons were made on a flat-top antenna only 50 ft. high and the efficient 240 vertical in span *B-D*. The flat top was 80 ft. in diameter with a central downlead extending to a 200-ft. diameter radial wire counterpoise. The fundamental wavelength was 288 meters and the resistance at 380 meters 5 ohms; of which at least 1 ohm is known to have been due to losses in the temporary tuning equipment used. Measurements were made with a graphic recording instrument on both day and night transmissions at 50, 120, and 170 miles. The results of these measurements indicate no marked difference in signal strength either at 50 or 120 miles; but at 170 miles the vertical was distinctly superior on both day and night transmissions. However, additional data supplied by independent observers indicate that propagation conditions, at the time of the 170-mile measurements, may have favored the vertical. Nevertheless, there is no question but that a difference was measured. It therefore appears either that the vertical was superior at 170 miles or that propagation conditions influence radiation from a high vertical in a different manner than the radiation from a low flat-top.

The use of different antenna forms may result in changes of energy concentration from the horizon to the zenith. If such changes occur they will alter the apparent antenna efficiencies based on the local field measurements. There is a possibility that such an effect may have been partially responsible for the shape of the curves of Fig. 6. The following discussion on this point is an extension of conclusions arrived at mathematically by various investigators.

From the voltage and current distribution along an ideal quarter-wave antenna it will be seen that the greatest electric charge is on the upper part while the magnetic field is most intense near the lower end. Considering the electro-magnetic field surrounding an element of antenna length near the base, all of the electric lines of force from the antenna will pass through the zone in a vertical direction; and the magnetic lines will be at a maximum and in horizontal planes. Bearing in mind that the direction of wave propagation is at right angles to both the electric and magnetic fields; and that the energy acting is proportional to the vector product of the intensities of the two fields: it will be apparent that the radiation due to this zone will have a maximum intensity and will be propagated in a horizontal direction. Taking zones surrounding successive elements of length higher and higher upon the antenna, the magnetic field becomes less and less and the average direction of the electric lines approaches more nearly the horizontal. Consequently, with increasing height the contribution to the total radiation of the zones surrounding the antenna elements becomes less and less and the direction of propagation approaches the zenith.

With a fixed operating wavelength, a decrease in antenna height from a quarter wavelength requires that a loading coil be used at the base; and, considering the voltage and current distributions on the shortened antenna, it appears that this is equivalent to winding the lower portion of a quarter-wave antenna—which contributes most to low-angle radiation—into a non-radiating coil. The part of the antenna contributing mostly to high-angle radiation still remains; and this may cause an upward tilt of the radiated wave, thus producing an apparent decrease in antenna efficiency when based on local field measurements

In the mathematical analysis of antenna characteristics given in "Electrical Oscillations and Electric Waves," by G. W. Pierce, it is assumed that a flat-top antenna consists substantially of a quarter-wave antenna with the upper part bent over in such a way that the voltage and current distributions remain unaltered. This, however, would represent only one limiting form of flat-top. The other limit would be reached if the dimensions of the flat top were expanded symmetrically about the vertical length so as to retain the same fundamental wavelength that existed with the quarter-wave antenna. The flat-top might then be assumed to be equivalent to replacing the upper part of the antenna by a circular conducting plate, in such a way as to retain the same field distribution about the remaining vertical length, as existed with the quarter-wave antenna. On this assumption, the only remaining part of the quarter-wave antenna is that which contributes mostly to horizontal radiation; and the use of such an antenna should result in concentration of energy close to the horizon. The characteristics of the "T" antenna used in securing the curve of Fig. 6 may have approximated those of the second limit closely enough to produce the effect of no apparent decrease in radiation efficiency with decreasing antenna height.

When antennas of low height are used, the losses in the ground system become an important factor in radiation efficiency. The question therefore arises as to the merits of an elevated counterpoise as compared with a buried wire ground. It is possible that, at broadcast wavelengths, the action of buried wires in providing a low-resistance path for the induced electric field is not very pronounced. The impedance of the wires themselves will be high due to their reactance at these frequencies while the impedance of the capacitive leakage path between wires will be low. A flow of current along the wires must be accompanied by a voltage drop and, as the potential of the uninsulated wires remains practically the same as that of the surrounding soil, it appears that the effectiveness of the wires is decidedly limited. On the other hand, an insulated counterpoise can assume a high potential which being opposite in sign to the potential of the antenna, facilitates the action of the counterpoise as a collector of lines of force. If the counterpoise is of sufficient height and dimensions, no appreciable earth currents should be present near the antenna; and due to the multiplicity of wires, the resistance of the counterpoise should be lower than that of the antenna conductor itself. Nevertheless, a counterpoise has obvious mechanical disadvantages as compared to a buried wire ground, so that its use should be conclusively justified by increased antenna efficiency. Such data as have been accumulated at the Developmental Station indicate that an insulated counterpoise does not increase the efficiency of a very high broadcast antenna by more than a small percentage. There is no question but that lower antenna resistance results from the use of a suitable counterpoise; but apparently a part of this is due to a reduction in radiation resistance. An example of extreme variations which may occur is shown by Fig. 7, the curves of which were based on data secured on the "T" antenna in span *D-F*. The measurements on the counterpoise

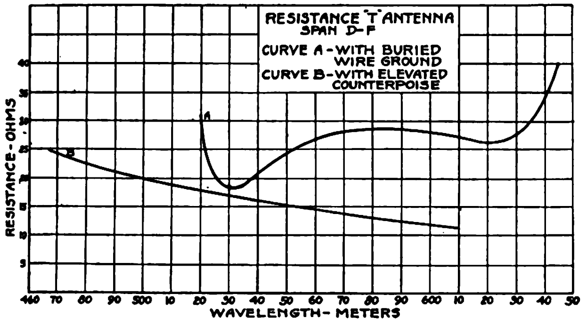


Fig. 7.

were made immediately after those with the wire ground, with no changes to the antenna itself. Therefore, the irregularities shown by the curve for the buried wire ground are definitely due to the use of this system.

Ground losses, far from being confined to the counterpoise or wire ground, extend to every part of the earth's surface reached by the radiated wave. Assuming the correctness of the ordinary radiation laws, the radiation field varies inversely as the distance from the antenna; and the induced field varies inversely as the square of the distance, being equal to the radiation field at a distance beyond the limits of the average ground system. For this reason, the losses in the induced and radiated fields overlap, thus making a clear division between power lost and power radiated practically impossible. Special precautions to reduce ground losses close to the antenna are advantageous simply because the induced field has comparatively a very high intensity in this region. For high efficiency as measured at distant points it is just as necessary to secure a favorable ground path for the radiated field; and this can be accomplished only by careful selection of the broadcast station site.

An indication of the extent of the ground losses occurring locally in the radiation field of an antenna can be obtained by field intensity measurements. For this purpose the measurements should be made along several radial lines from the antenna, and the decrease in field intensity due to the normal spreading of the waves taken care of by multiplying each measured value by the distance of the measurement point from the antenna. The degree to which an average curve (of these "intensity by distance" values plotted against distance) varies from a horizontal line should be indicative of the decrease in field intensity due to ground absorption.

Measurements of this kind were made on two broadcast stations which were equipped with almost identical antennas and operated at not greatly different wavelengths, but whose locations were very different as far as soil and topography are concerned. The curves obtained from these measurements are shown by Fig. 8, and indicate that station No. 2 would require four times the power input of station No. 1 to produce the same average signal strength at 20 km. Assuming the same average absorption beyond 20 km for both stations, the advantage of the location of station No. 1 will be apparent. The differences indicated by these curves, while probably extreme, show the importance of securing the best possible antenna location within the permissible limits. Even where the choice of location is decidedly restricted, measurements made with portable equipment prior to final selection of the site may be helpful in avoiding the vicinity of small highly absorbing areas. The power lost in such an area will depend on the percentage of total radiated power acting through the area; and if the area is at a considerable distance from the antenna, it will occupy an arc which is only a small percentage of the radiation circle and consequently cause but little loss.

The effectiveness of increasing the power output of a broadcast station will be reduced by ground absorption of the radiated field. The object of increased power is greater broadcast range; and for the word "range" to have a meaning we must assume the existence of some minimum signal necessary for normal broadcast reception. With a circular radiation pattern and uniform propagation conditions, the range of the station would be the distance from the antenna to a point where this minimum signal was obtained; and the area served would be proportional to the square of this distance. Two conditions act to reduce the radiated field intensity: the normal spreading of the waves by which doubling the circumference (or radius) of the radiation circle cuts the intensity of the field in half; and absorption, acting according to the inverse compound interest law, by which a certain percentage of the existing field is lost for each unit of distance traveled. Now, increasing the power output of a station to four times its previous value will double the field intensity at the point of first minimum signal level regardless of absorption;

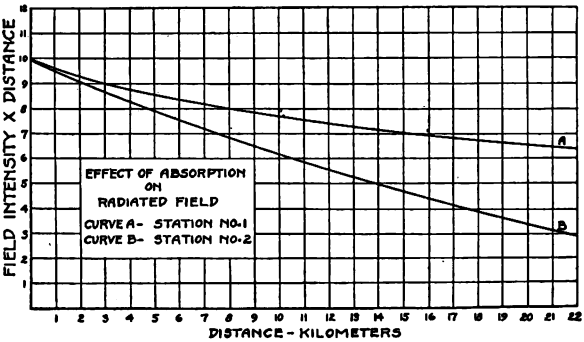
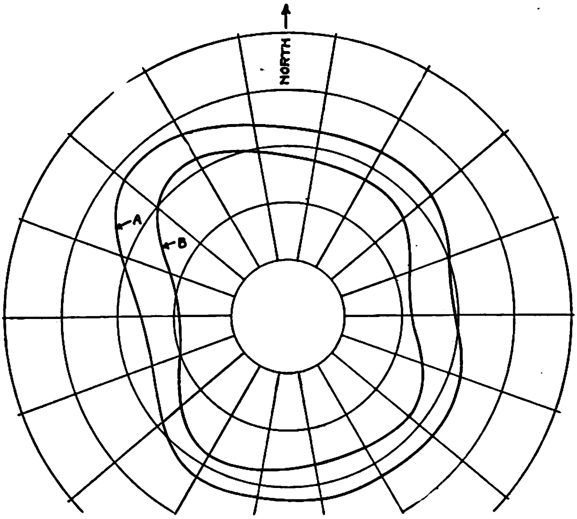


Fig. 8.

effects of absorption. Therefore, the shape of the calculated or measured field intensity pattern will be different from that of the minimum signal level curve. The field pattern shows variations of intensity at a constant distance from the antenna, while the minimum signal curve shows variations in distance for a constant field intensity; and the percentage of absorption varies with the distance from the antenna and not with field intensity. The result, with uniform propagation, will be that the minimum signal curve as compared with the field pattern will show the high points reduced and the depressions partly filled out. That is, the minimum signal curve will approach more nearly a circular form than indicated by the field intensity pattern. This effect is probably responsible to a large extent for the apparent "healing" of distorted signal level curves as the distance from the antenna is increased.



FIELD PATTERN COMPARISON VERTICAL & FLAT TOP ANTENNAS

PATTERN A- VERTICAL ANTENNA 240 FT. LG.
PATTERN B- "T" ANTENNA. HORIZONTAL LENGTH 240 FT.
VERTICAL LENGTH 80 FT.

Fig. 10.

As an example of directive radiation accidentally obtained, the general form of field pattern on the regular WGY antenna is shown by Fig. 8. This antenna was previously described as consisting of two inverted "L" antennas placed end to end. The separation distance of the downleads, 200 ft., is approximately one-sixth of a wavelength; and the space-phase difference brought about by this separation, together with a similar possible time phase relation, is sufficient to account for the elliptical form of the pattern. The usual flat-top lengths of inverted "L" and "T" antennas used for broadcasting are not sufficient to produce directive radiation. This is indicated by the almost exact shape of the two field patterns shown by Fig. 9, one of which was made on a vertical antenna 240 ft. long and the other on an 80-ft. "T" antenna with a 240-ft. horizontal section. Such directivity as does exist in these patterns is due to the action of the 300-ft. supporting towers.

The ultimate conclusion from such investigations as those discussed in this paper must be that each broadcast antenna presents a separate engineering problem requiring individual treatment. A great advance in establishing a sound engineering basis for the attack of such problems would be made by comparing the broadcast range of a number of existing stations. Work of this nature has in the past been hindered by the lack of suitable measuring instruments and by the absence of an authoritative field intensity standard. Precision instruments are now available capable of measuring the intensity of the radiated field under full power conditions, from within a few wavelengths of the antenna to the limits of the station range. Calibration methods have also been extensively studied, with the result that measured intensities are considered to be close to the true values in microvolts per meter.

A standard method of instrument calibration giving a fairly close approximation to microvolts per meter is a most urgent requirement. Scientific accuracy is less important at the present time than a general basis for comparison of the broadcast ranges of existing stations. Of course, the distance of minimum signal level is obscured by fading, but we can take the daylight range as a performance basis; and assume, until we learn more about fading and radiation generally, that an antenna which is good in the daytime is also good at night. By considering each station as a separate problem in radiation we should be able to do much towards increasing the effectiveness of the stations. The result would be that we would have less broadcasting, in the literal sense of the word, and more of what the industry seems bound to become, a thoroughly efficient public service.

DEVELOPMENT OF RADIO AIDS TO AIR NAVIGATION*

By

J. H. Dellinger and Haraden Pratt

(Bureau of Standards, Washington, D. C.)

Summary—*The development of aircraft has reached a stage where further progress depends upon the conquest of adverse meteorological conditions. The most promising avenue to such mastery of the elements is through adaptations of radio. Intensive work extending over two years has resulted in the development and practical demonstration of a complete set of radio aids to flying on the civil airways of the United States. This development was carried on by the Bureau of Standards for the Aeronautics Branch of the Department of Commerce.*

The radio aids, which will now be installed on the principal airways, comprise a radio beacon system and telephone service from ground to aircraft. The required radio equipment on the airplanes is reduced to a short pole antenna and a simple receiving set weighing a few pounds, including a visual indicator which tells the pilot whether he is on the course or how far off. All of the expensive and powerful apparatus necessary for the system is on the ground, maintained by the Government.

The radio beacons operate in the frequency band 285 to 315 kc and the telephone stations in the band 315 to 350 kc. These are allocated to air service by the 1927 International Radio Convention. For the present the beacons are adjusted to the frequency of 290 kc and the telephone stations to 333 kc.

The directive radio beacon is a special kind of radio station, usually located at an airport, just off the landing field. Instead of having a single antenna like an ordinary radio station, it has two loop antennas at an angle with each other. Each of these emits a set of waves which is directive, i.e., it is stronger in one direction than others. When an airplane flies along the line exactly equidistant from the two beams of radio waves, it receives signals of equal intensity from the two. If the airplane gets off this line it receives a stronger signal from one than the other.

The indicator connected to the receiving set on the airplane shows when the signals from the two beams are received with equal intensity, by means of two vibrating reeds which are tuned to different modulating frequencies used on the two antennas at the directive radio beacon station. When the beacon signal is received the two reeds vibrate. The tips of these reeds are white in a dark background so that when vibrating they appear as a vertical white line. The reed on the pilot's right is tuned to a frequency of 65 cycles and the one on the left to 85 cycles. It is only necessary for the pilot to watch the two white lines produced by the vibrating reeds. If they are equal in length, he is on his correct course. If the one on his right becomes longer than the other, the airplane has drifted off the course to the right (into the region where there is

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Publication Approved by the Director of the Bureau of Standards of the Department of Commerce, Washington, D. C.

more of the 65 cycles). If he drifts off the course to the left, the white line on the left becomes longer.

Successful flights have been made up to 135 miles, in fog and over hazardous mountain terrain. This distance is more than enough to demonstrate the success of the system, as it is contemplated that the directive radio beacon stations will be placed not over 200 miles apart. Beacons placed at such distances, with a straight airway between them, will be supplemented by small marker beacons at intervals along the route. A characteristic signal from a marker beacon will show on the visual indicator aboard the airplanes what point is being flown over.

The whole receiving system comprises a small indicator unit on the instrument board weighing one pound, a receiving set weighing less than 15 pounds, and a 15-pound battery. The receiving system is very little affected by interference, including engine ignition interference, which has hitherto been the bar to satisfactory use of radio on airplanes.

INTRODUCTION

THE urge for dependable instrumentalities of aerial navigation has become progressively more intense: a natural consequence of the development of more reliable aircraft, the successful accomplishment of pioneer long-distance flying, and the commercial operation of scheduled air services. The operation of aircraft is peculiarly subject to uncontrollable meteorological conditions. The very navigational means which such conditions demand have hitherto been lacking.

The period preceding the World War saw aeronautics develop out of the stage of experimentation into the stage of fact and reality. The war provided a field of use which stimulated intensive development in everything aeronautic, creating a foundation of experience upon which this new instrumentality of transportation has since grown. During the war the use of radio as an adjunct to military flying received some attention, and both telegraph and telephone sets were successfully used for moderate distance communication.

In Europe, aeronautic post-war development took the form of subsidized air passenger lines operating on schedule. The problem of maintaining schedules during fog and bad weather very soon required the use of radio as a navigational aid. The form in which radio has evolved, to assist bad-weather flying in Europe, is a two-way radio telephone (or telegraph) communication service between ground and aircraft, supplemented by a system of position finding by means of direction-finder stations on the ground. Upon radio request from an aircraft, two or more ground direction-finder stations locate the aircraft by triangulation and send the information back. The operation is similar to that of the radio compass service rendered by our Navy Department to merchant ships entering American ports.

In our country, aeronautic post-war development largely took the form of air mail service, operated first by the Government and subsequently by private contractors. The first successful experiment in regular cross-country night flying was instituted by our Post Office Department between Chicago and Salt Lake City in 1924. Besides establishing lighted airways, the Department also provided advance weather service by radio telegraph between landing fields.

These substantial advances in aerial transportation brought out more strongly than ever before the need for better navigation facilities, as upon these depended the betterment of flying schedule reliability and the reduction of the hazards associated with unfavorable weather.

Recognizing this need, under the Air Commerce Act of 1926 establishing commercial airways, Congress provided funds to develop and maintain aids to navigation on the airways. Accordingly, shortly after its formation in July, 1926, the Aeronautics Branch of the Department of Commerce began a program of establishing aids to navigation on the nation's airways. The aids determined upon were: upper air weather information, airways lighting for night flying, and radio aids. It was recognized that radio offered important advantages in the way of communication between the ground and air and also direct navigational assistance. As there had been little or no experience in the use of radio under the conditions of flight on the civil airways in the U. S. (the Air Mail Service did not employ radio on its airplanes), it was not immediately evident what form these radio aids should take.

A research division of the Aeronautics Branch was organized in the Bureau of Standards to undertake the necessary experimentation and development in this field. It is the development work so far done under the Act that this paper describes.

EARLY WORK

As already mentioned, radio communication between aircraft and ground received some development during the war. Such communication is a powerful aid to air navigation, since in conditions of storms or poor visibility the pilot can be informed of conditions along his route or told where a landing can be most safely effected. The first attempts to develop a radio device as an actual navigation instrument involved the direction finder, which had been developed and used with great success in marine navigation. It was not possible to duplicate this success in air navigation. Engine ignition system interferences, limited space, excessive noise, and preoccupation of the pilot diminish the probability of taking useful bearings on airplanes with direction-finding coils. Even the use of schemes[1] whereby a bearing could be taken by comparing head telephone signals of equal intensity do not escape these disadvantages. There have been instances[2] of successful use of direction finders on airplanes, but the disadvantages mentioned are inherent.

The reverse procedure[3] (use of direction finder on the ground) is employed successfully abroad, as previously mentioned. However, only aircraft having complete transmitting and receiving sets can receive aid from this system. This eliminates the small airplane. Furthermore, errors in the bearings taken are apt to result with this system owing to the inclination of the airplane transmitting antenna.[4]

The next effort to develop a radio navigational device was the development by the Bureau of Standards of a special type of radio beacon in 1921. Working on the air navigation problem at the request of the Army Air Service, the Bureau devised the crossed-coil or double-beam directive type of radio beacon.[5] This beacon consists essentially of two separate coil antennas, set at an angle with each other, which send signals at alternate intervals. Equality of signal strength from the two antennas is obtained along a line bisecting the angle between the planes of the two antennas. The advantage of the system is that its use requires nothing more than a radio receiving set on the airplane.

In the early work a switch was used to throw the radio-frequency power from one antenna to the other.[5] Tests made at

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[1] "Direction Finders for Aircraft," J. Robinson; *Wireless World*, **7**, p. 475; 1919.

[2] "Use of Radio in Airplanes," P. Franck; *L'Onde Electrique*, **2**, p. 200; 1923; **7**, p. 109; March, 1928.

[3] "Wireless Bearings from the Air." *Wireless World and Radio Review*, **18**, pp. 866–868; June 30, 1926.

[4] "Stationary and Rotating Equisignal Beacons," W. H. Murphy and L. M. Wolfe. *Journal* Society Automotive Engineers, **19**, p. 209; September, 1926.

[5] "A Directive Type of Radio Beacon and its Application to Navigation." F. H. Engel and F. W. Dunmore, Bureau of Standards Scientific Paper No. 480, 1923.

Washington on the ground and on ships showed that a course was effectively marked out, permitting navigation without aid of landmarks, compass, or other navigational device. The apparatus was next set up at Dayton, Ohio, and tests made in the air. The method was successful in airplane flights, and had the important advantages that it required no special apparatus to be carried on the airplane and that no error was introduced by wind drift.

In the following four years, the Army engineers at Dayton, Ohio, developed the beacon further[4]; in particular, they devised a signal-switching arrangement such that the signals from the two antennas merge into a steady dash when on the course, giving an added criterion besides that of equal signal intensity to enable the observer to tell whether he is on or off the course. They also introduced a goniometer, or mutual inductance device, to permit orienting the course in any desired direction without moving the antennas.

RECENT WORK

Under the stimulus of the Air Commerce Act, the Bureau of Standards again undertook this field of investigation in 1926, this time for the newly formed Aeronautics Branch of the Department of Commerce. As a first step, a conference was called June 22, 1926, to secure information and advice from various organizations which had experience bearing on these problems. As a result of its recommendations the development of radio aids for the airways of this country was immediately begun, the principal features to be radio telephony from ground to aircraft, and a radio beacon system.[6] One of the major considerations in narrowing the choice of radio aids to these two was the idea of minimizing the apparatus which would have to be carried aboard the airplanes. The system adopted requires only a simple receiving set to be carried by an airplane. With this set it receives radio telephone information on weather conditions, etc., and also the radio beacon signals to guide it along its course. All of the complicated and expensive apparatus is on the ground at the transmitting end, to be maintained by the Government.

During 1926 and 1927 the Bureau established two experimental radio telephony and beacon stations to carry on its work, one at College Park, Md., and one at Bellefonte, Pa. The latter was chosen so that practical radio test flights and service flights could be undertaken over a difficult sector of the transcontinental airway.

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[6] "Applications of Radio in Air Navigation," J. H. Dellinger. *Engineers and Engineering,* **43,** p. 301; Nov. 1926; *Mechanical Engineering,* **49,** p. 29; Jan., 1927.

INTERLOCKING TYPE OF DIRECTIVE BEACON

A simplified circuit arrangement of the beacon is shown in Fig. 1. A 50-watt master oscillator with two 250-watt amplifier tubes supplies power to two loop antennas. A 500-cycle generator supplies plate power. With this arrangement coil antennas of a type capable of rotation are used to permit orientation of the course.

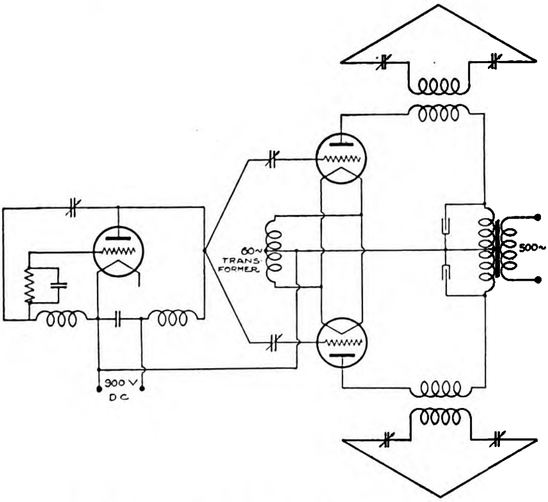


Fig. 1.—Diagram of Elementary Directive Beacon Circuit Employing Rotating Loop Antennas.

While some of the first beacon set-ups employed quenched-spark sets to provide the radio-frequency power, the College Park beacon started from the first with vacuum-tube equipment.

To permit orientation of the course in any direction when large fixed antennas are used, the circuit arrangement shown in Fig. 2 was set up at College Park. This is substantially the arrangement developed by the Army engineers. A four-coil goniometer was interposed between the amplifier plate circuits and the antennas, so that the currents in these antennas due to the driving voltage of one primary goniometer coil would create a resultant field corresponding to that which would be produced by an imaginary or phantom loop rotating with the primary coil. Two such phantom loops, rotating as the two primary goniometer coils rotate together, set up the course or equisignal zone and allow it to be oriented at any angle without any variation in the field intensity produced.

Fig. 2 also shows the circuit arrangements for the sending of the interlocking signal. As the cams rotate slowly, amplifier tube 1 is excited with the Morse letter *A* and amplifier tube 2 with the letter *N*. The two are so interlocked that if the outputs of both amplifiers were fed into a common circuit only a long dash would result. It is to be noted that tubes 1 and 2 never receive any grid excitation simultaneously. An observer located on a line bisecting the angle between the two phantom loops, receiving signals of equal amplitude from each loop, would

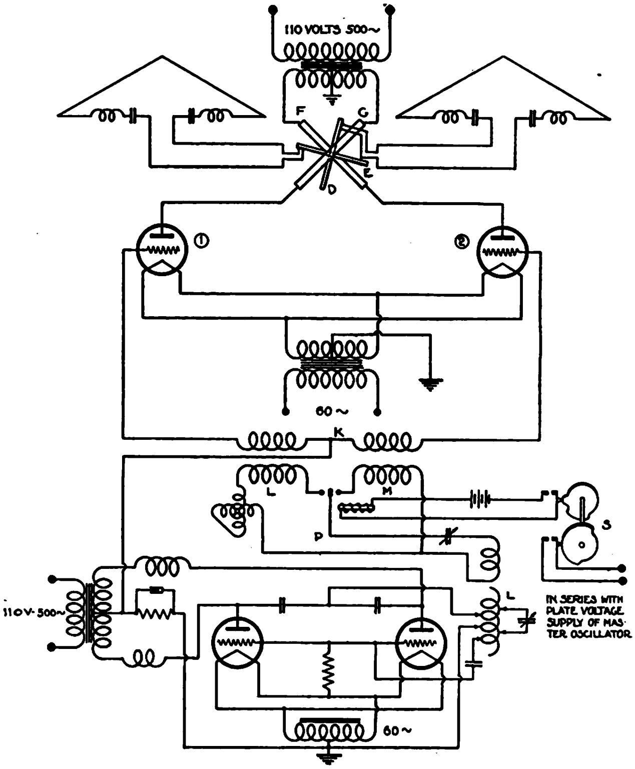


Fig. 2.—Diagram of Interlocking Signal Type of Directive Beacon Circuit Equipped with Fixed Loop Antennas and Goniometer.

therefore hear only long dashes. Were he not on such a line, then he would hear a preponderance of either letter *A* or *N* depending upon the distance away from the line.

Fig. 3 shows a diagrammatic representation of the interlocking signal for different angular positions around the beacon. As might be concluded from inspection of this diagram, some training and skill are necessary to determine direction sharply by listening to these signals.

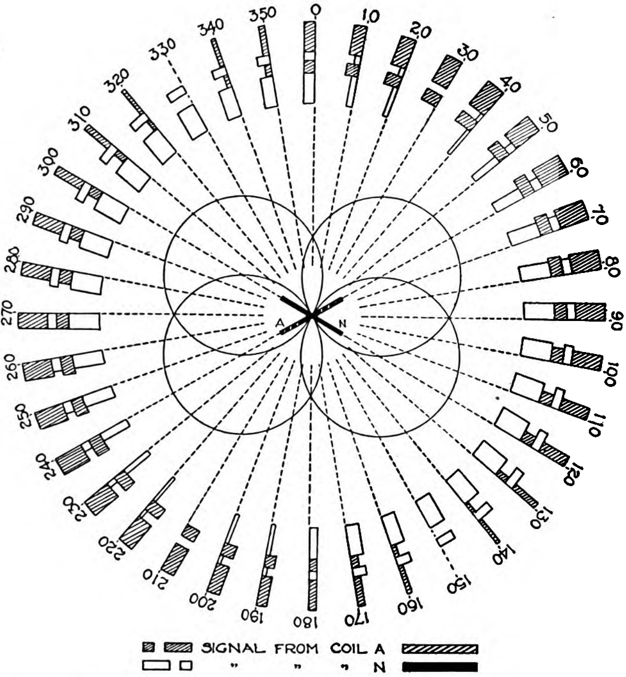
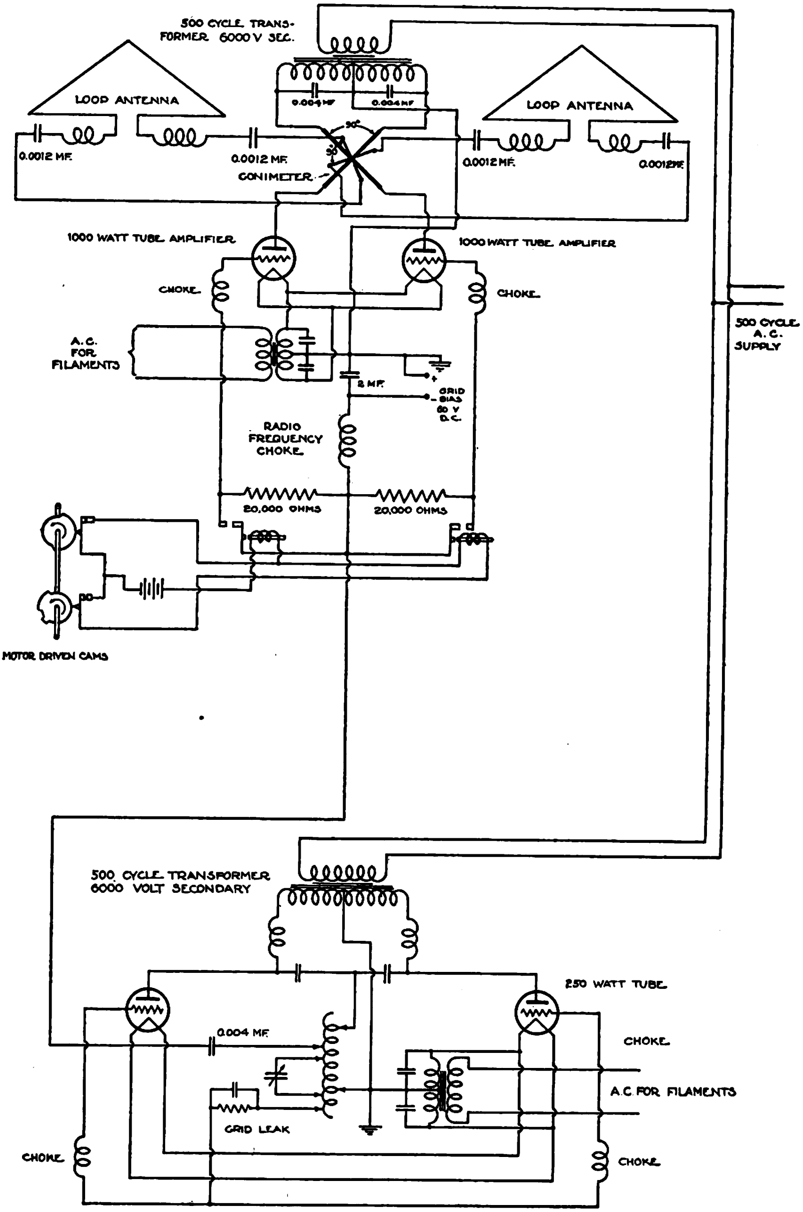


Fig. 3.—Interlocking Characteristic of Equisignal Radio Beacon, Showing Change in Relative Intensity of the Two Signals Received on an Airdraft Flying around the Beacon.

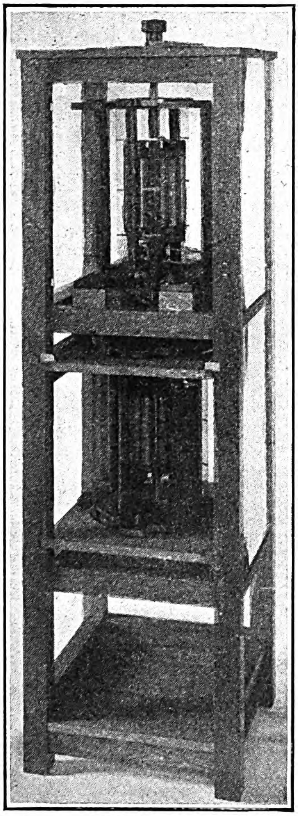
The diagram of Fig. 4 shows further refinements and simplifications resulting in a circuit which has been in use for practical test flying for nearly a year. Small relays serve to provide the interlocking through control of grid current. The master oscillator uses one 250-watt tube and the amplifiers two 1000-watt tubes. The type of goniometer used is shown in Fig. 5. It has two rotors each wound with 8 turns of No. 12 bare copper wire and two stator coils each wound with 32 turns of insulated wire.



MASTER OSCILLATOR OF SUFFICIENT SIZE TO SUPPLY VOLTAGE TO EXCITE AMPLIFIER TUBE GRIDS.

Fig. 4.—Improved Directive Beacon Circuit Diagram Showing Connections of Interlocking Mechanism.

The stator coils are the primary input coils from the amplifier tube plate circuits. It is because this gives a more simple mechanical design that the secondary coils are the ones capable of rotation. With this construction, therefore, the phantom loops revolve in the opposite direction to the revolution of the rotor coils, and care must be exercised to engrave and mount the indicating dial to take that fact into account.



F g. 5.—Double Rotor Type of Goniometer Used with Modulation Type Directive Beacon.

A number of disadvantages are inherent in the aural interlocking beacon system as described, and the Bureau undertook to investigate these and seek measures for improvement. The obvious advantage of having a visual indicating instrument rather than using the aural (telephone receiver) method made the development of a visual system the first objective. Another objective was the reduction or elimination of apparent errors in the observed course due to slanting airplane antennas and other causes.

Several methods of operating a visual device for a directive beacon system have been tried. The Army Air Corps worked out a sequence relay arrangement whereby the letter *A* of the interlocking signal would light a red lamp on the airplane instrument board, the letter *N* a green lamp, and the interlocked dash a white light. This system has not been found practical because of the complications of the special apparatus and its weight, and also the disadvantage that it indicated only whether the airplane was on or off the correct course, giving no information as to how far off it might be.

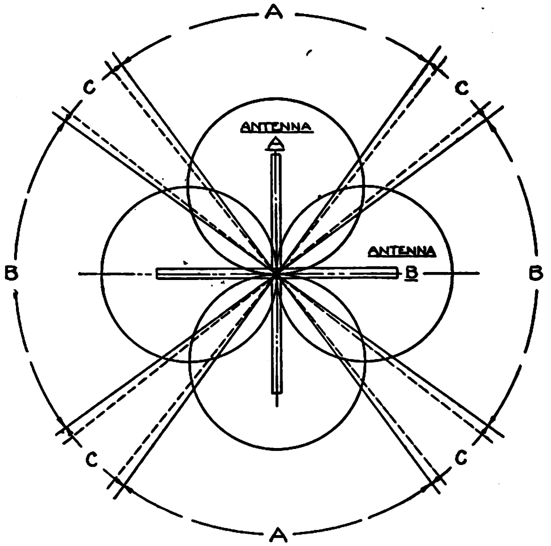


Fig. 6.—Field Intensity Distribution Diagram for Early Type of Modulation Beacon System.

Modulation Type Beacon

The Bureau attacked the problem of visual indication from another angle. It was determined to try the possibilities of energizing both antennas simultaneously and modulating each by a different audio frequency. Means for separately selecting these modulated frequencies in the receiving set would permit the operation of some indicating device.

As a first step in this development, a method was tried which required only a single source of audio-frequency modulation. Reference to Fig. 4 shows that if both amplifier tube grids receive excitation from the master oscillator simultaneously, power can be supplied to both loop antennas together. Owing to the method of supplying plate voltage whereby plate current flows only when the audio-frequency alternating supply voltage is positive, radio-frequency current flows in only one antenna at any given time.

An observer, therefore, exploring the field around a beacon station so connected and excited from a 500-cycle alternator, will encounter, as indicated in Fig. 6, two wide zones *A* where signals of 500-cycle pitch from loop antenna *A* will be heard. Similar signals will be heard in two other wide zones *B* from loop antenna *B*. In four other narrow zones *C*, signals of 1000-cycle pitch will be heard, due to the combined presence of approximately equal signals from both loop antennas. Since the audio-frequency modulation of one antenna lags one-half cycle behind that of the other, the two received together produce the 1000-cycle pitch. If the observer moves from a zone *C* to either zone *A* or *B*, he finds a location where the amplitude of the received 500-cycle current equals that of the 1000-cycle. The eight dotted lines indicate the location of the eight places where such equality exists.

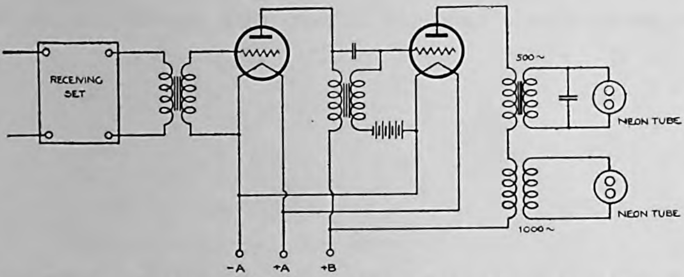


Fig. 7.—Circuit Diagram for Visual Indicator Using Glow Lamps.

It was thought that if a suitable visual device could be developed to indicate when equality of the 500- and 1000-cycle currents was present, then these dotted lines would constitute radio courses. Such a beacon transmitter contained the very desirable feature of simplicity, particularly as with a push-pull connection of two tubes in the master-oscillator circuit the same alternator could be used for all tubes, grid excitation voltage in this way being present whenever either amplifier tube plate is positive.

Two visual methods were attempted. Fig. 7 shows a neon lamp device. The output of a receiving set was fed into two selective circuits containing the lamps. Both lamps lighted indicated the position of the course location. One lamp lighted meant off course on the side corresponding to the frequency used to light it. This method was abandoned after several trials because of the need for too many amplifiers to operate the lamps, nor could the distance deviated from the course be told as the lamps were either lighted full or not at all.

Fig. 8 shows the other visual scheme tried. Here the selected audio-frequency currents are rectified and the direct-current outputs connected opposed to a center zero d.c. microammeter. The needle stands balanced at zero with equality of the 500- and 1000-cycle currents, with a swing in the appropriate direction resulting from "off course" positions. The observer here could estimate his distance off course, but a new disadvantage arose. If all is well and the airplane is flying on the course with balanced needle at zero, the beacon might accidentally stop without the pilot becoming aware of it. This actually occurred twice during test flights near College Park.

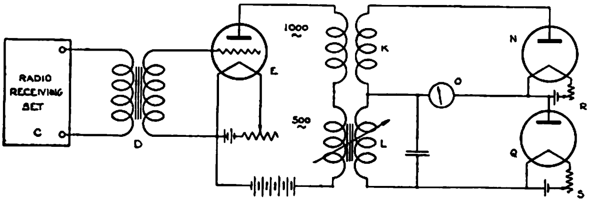


Fig. 8.—Circuit Diagram for Visual Indicator Using Zero Center Microammeter.

Although much work was done on this visual system, it was finally laid aside as impractical because of serious inherent defects. Unduly special receiving circuits were needed. The course could be shifted while flying by changing the radio-frequency tuning, thereby changing the relative magnitudes of the side-band amplitudes corresponding to the 500- and 1000-cycle currents. Frequent rebalance of the rectified currents through the microammeter was necessary. And last but not least, eight courses around a beacon, with pairs of them only a few degrees apart, proved to be very confusing to the pilots who had opportunity to try it, constituting an arrangement which might have an element of danger.

It was therefore decided to try a system in which two independent modulation frequencies are supplied in the beacon transmitter. This system results in four instead of eight courses, as shown in Fig. 9. To eliminate the wide sidebands of the previous system, low modulation frequencies seemed desirable. It then became possible to use a very simple vibrating steel reed as an indicator. On the airplane the receiving set output is fed to the indicator consisting of small electromagnets between which two vibrating polarized steel reeds are mounted. The reeds are tuned mechanically to the frequencies of the beacon modulation voltages.

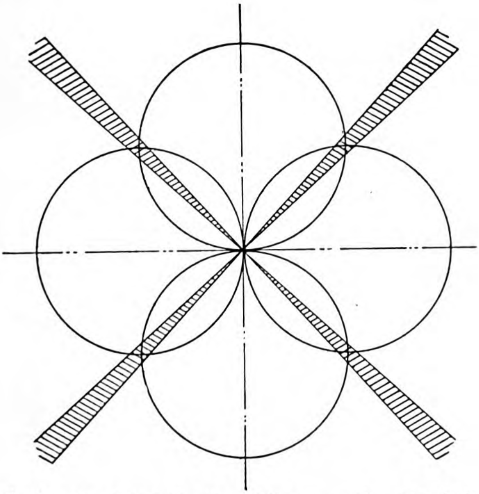


Fig. 9.—Field Intensity Distribution Diagram for Modulation Type of Beacon.

Simplicity of operation has resulted. An ordinary receiving set, with the small reed unit which weighs less than an ordinary pair of head telephones, constitutes the airplane radio equipment. The reeds vibrate with equal amplitude when on the course. When off the course, the observer can estimate the distance he is off by noting the relative reed amplitudes. An advantage of the system is that if the beacon ceases to function the observer is aware of it, whether he is on or off course. A still more important advantage is that, owing to the sharp mechanical tuning of the reeds, they are immune to interference that would entirely ruin aural reception, unless the interference is of such strength as to overload completely the receiving set.

BEACON TRANSMITTING APPARATUS

The first practical attempt to produce workable signals was carried out with a master oscillator operating with a direct-current plate supply. One of the amplifiers was supplied with plate power from a transformer connected to the commercial 60-cycle supply line, and the other amplifier supplied with 86 cycles obtained from a small alternator. Two reeds tuned to these frequencies were made up and full scale deflections were obtained at a distance of ten miles with a suitable receiving set. Upon trial in flight an observer equipped with reeds tuned to these modulating frequencies experienced no difficulty in guiding the airplane by giving instructions to the pilot.

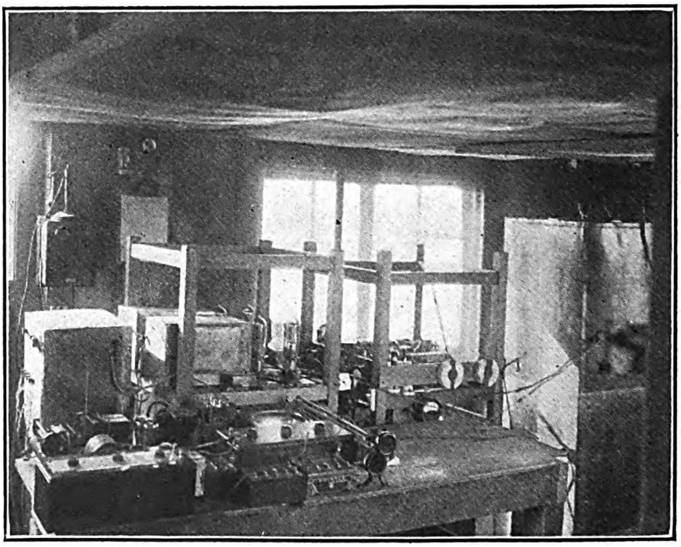


Fig. 10.—Temporary Beacon Apparatus for Modulation System, Installed at College Park, Md.

It was observed when making a field intensity survey of signals from this beacon, that symmetrical figure-eight patterns were not secured when the stator coils were set at right anles to each other. It was found necessary to set the stator coils that there would be no inductive coupling between them. Some of these effects appeared due to couplings between different parts of the system. It was also found that a more steady source for the low-frequency currents would be needed. The commercial power supply frequency varies in Washington as much as ±0.15 cycle while the reeds required steadiness within ±0.05 cycle.

It was accordingly decided to secure the low frequencies from two vacuum-tube driven tuning forks and then modulate the 290-kc carrier in the two respective amplifier branches from these using either grid or plate modulating methods. The system was therefore set up in this way using d.c. plate power supply throughout, though at first vacuum-tube oscillators were employed instead of tuning forks. Their frequencies were held sufficiently steady by hand-operated trimming condensers so they could be used for experimental purposes. Fig. 10 shows the first temporary setup of apparatus at College Park. Fig. 11 shows the circuit arrangement. The master oscillator employs a 7.5-watt tube. This supplies grid voltage for both of the first tubes of the two sets of amplifiers, which are also of 7.5-watt size. Following this is a stage consisting of two 50-watt tubes in parallel followed by a single 1000-watt tube. Voltage from the low-frequency oscillators marked *A* and *B* in Fig. 11, which also use a 7.5-watt tube each, is applied in each case to the grids of two 7.5-watt modulator tubes connected to the plate of the 7.5-watt radio-frequency amplifier tube in the usual Heising fashion. A goniometer of different design was also introduced, having two sets of rotor coils in series, each set being in the field of one of the stator coils. The purpose of this construction was to separate the stator coils and reduce coupling between them.

Upon investigation many stray capacitive couplings were found in the system. For example a 10 to 15 per cent capacitive coupling was found to exist between the stator coils due to their position with respect to each other. A like amount was found between the stator coils due to the couplings between them and the common rotor system which, while it did not manifest itself as a coupling between the stators, resulted in capacity currents flowing out through the rotor coils into the horizontal portions of the loop antennas. These currents, being of different phase from those normally flowing, introduced errors and dissymmetry. These couplings were largely overcome by placing copper

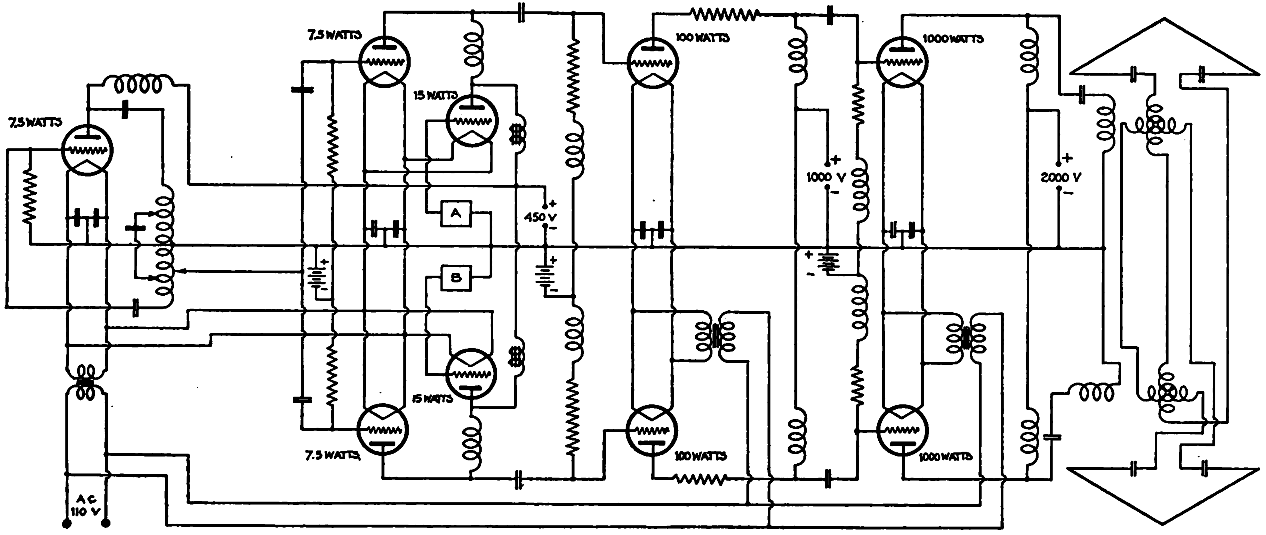


Fig. 11.—Circuit Diagram of Double Modulation Type of Beacon. *A* and *B* are tuning-fork generators of modulation frequencies.

screens between and around the stator coils, and grounding both rotor systems at their center point so that capacity currents leave the system by that return instead of via the antenna wires. It was also found necessary to shield carefully all the amplifier stages as well as the master oscillator, and introduce choke coils with bypass condensers at strategic points so as to prevent cross couplings between the two amplifier systems.

Some remarks concerning the field patterns produced by this system may be of interest. In vector diagram, Fig. 12

E_{S_1} is voltage across upper stator coil S_1 of goniometer
E_{S_2} is voltage across lower stator coil S_2 of goniometer

$$\left.\begin{array}{l}E_{R_1'}\text{ is voltage in upper rotor coil }R_1'\text{ of goniometer}\\ E_{R_1''}\text{ is voltage in lower rotor coil }R_1''\text{ of goniometer}\end{array}\right\}\begin{array}{l}\text{connected}\\ \text{in series}\\ \text{with an-}\\ \text{tenna 1.}\end{array}$$

$$\left.\begin{array}{l}E_{R_2'}\text{ is voltage in upper rotor coil }R_2'\text{ of goniometer}\\ E_{R_2''}\text{ is voltage in lower rotor coil }R_2''\text{ of goniometer}\end{array}\right\}\begin{array}{l}\text{connected}\\ \text{in series}\\ \text{with an-}\\ \text{tenna 2.}\end{array}$$

S_1 and S_2 are at right angles to each other
R_1' and R_1'' are in the same plane.
R_2' and R_2'' are in a plane at right angles to rotors R_1.
P is the position of the observer considered to be in the horizontal plane containing the radio station.

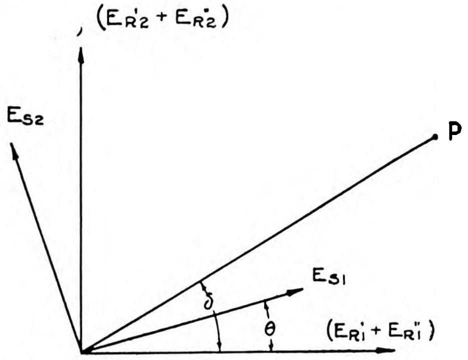


Fig. 12.—Vector Diagram of Goniometer and Antenna Currents.

For the sake of clarity, since the rotor currents are the same as the antenna currents, and the antennas are fixed in space, this treatment can be simplified by considering the stator coils as the ones rotated. This has not been done with the equipment because of mechanical reasons, a better structure resulting when the antenna coils are the ones capable of rotation.

Whenever observer P is in such a location that he obtained equal field intensities from the two phantom beacon antennas, and these intensities become unequal as the observer moves away transversely in either direction from the place of equality, then a radio course can be said to exist.

The field intensity at P will be

$$E_p = K[(E_{R_1'}+E_{R_1''})\cos\delta + (E_{R_2'}+E_{R_2''})\sin\delta] \tag{1}$$

The rotor voltages will be

$$E_{R_1'} = AE_{S_1}\cos\theta$$
$$E_{R_1''} = AE_{S_2}\cos(90+\theta)$$
$$E_{R_2'} = AE_{S_1}\sin\theta$$
$$E_{R_2''} = AE_{S_2}\sin(\theta+90)$$

The antenna voltages will be

$$E_{R_1'}+E_{R_1''} = AE_{S_1}\cos\theta - AE_{S_2}\sin\theta \tag{2}$$
$$E_{R_2'}+E_{R_2''} = AE_{S_1}\sin\theta + AE_{S_2}\cos\theta \tag{3}$$

where K and A are constants.

Combining equations (1), (2), and (3) and simplifying

$$E_p = AK[E_{S_1}\cos(\theta-\delta) - E_{S_2}\sin(\theta-\delta)] \tag{4}$$

This shows that for any fixed position of the coils, the field pattern will be two cosine curves at right angles to each other, as polar functions of angle δ.

The field intensity observed at a fixed point P varies in the same manner as the coils are rotated. If no modulation is impressed on either amplifier system, and taking

$$E_{S_1} = E_{S_2} = E_S$$
$$E_p = AKE_S\sqrt{2}\sin\left(\theta-\delta-\frac{\pi}{4}\right) \tag{5}$$

The existence of such a single figure-of-eight pattern which rotates when the coils are turned was verified by observation.

Of interest is the situation when one rotor coil is reversed. The vector diagram of Fig. 13 applies. Here equation (3) must be written

$$E_{R_2'} - E_{R_2''} = AE_{S_1}\sin\theta - AE_{S_2}\cos\theta \tag{6}$$

and equation (4) becomes

$$E_p = AK[E_{S_1} \cos (\theta - \delta)] - E_{S_2} \sin (\theta + \delta) \tag{7}$$

In this case when the coils are rotated the cosine and sine field intensity patterns revolve in opposite directions, and the course cannot be oriented where desired. It lies on the same axis at all times. This relation can be written in another form taking the case of no modulation, with

$$E_{S_1} = E_{S_2} = E_S$$

$$E_p = AKE_S(\cos\theta \cos\delta + \sin\theta \sin\delta - \sin\theta \cos\delta - \cos\theta \sin\delta)$$

$$= E_p = 2AKE_S \cos\left(\theta + \frac{\pi}{4}\right) \sin\left(\delta + \frac{\pi}{4}\right) \tag{8}$$

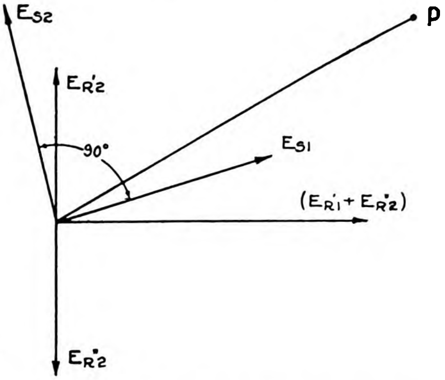


Fig. 13.—Vector Diagram of Goniometer and Antenna Currents with One Secondary Coil Reversed.

Here it can be seen that the only effect of changing the angular position of the coils, which means a change in θ, is to control the amplitude of the sine curve field intensity pattern. When $\theta = \frac{\pi}{4}$ or $\frac{5\pi}{4}$ no external field exists as E_p is zero. At angles $\theta = \frac{3\pi}{4}$ and $\frac{7\pi}{4}$ the two components of the field are in phase and no course exists in the sense as defined above.

Care must therefore be taken when connecting up an arrangement of this kind, not to get one rotor coil reversed. This can be very easily checked by noting the antenna currents. When correctly connected the antenna currents, resulting from the voltages expressed by equations (2) and (3), are such that when there is a rise in one, a fall occurs in the other when the goniometer is rotated. If incorrectly connected the currents are as expressed in equations (2) and (6) and rise and fall together as the goniometer is turned.

So far we have assumed resonance in the antennas with voltages and currents in phase. This has permitted the writing of a direct relation between field intensity and stator coil voltage. Another matter of interest, therefore, is what occurs when these antenna currents are out of phase. Such a condition can occur from several causes, but the most likely one from a practical standpoint is that due to the antenna circuits getting out of tune.

The loop antennas at College Park operate at 290 kc and are tuned with condensers of 0.0006 μf in series. The total loop resistance is about 8 ohms. If the capacity of one loop were increased and that of the other decreased a like amount, reducing the loop currents to $\frac{1}{\sqrt{2}}$ (their normal values) then these currents would be in quadrature. This would result in a rotating field and the field intensity would be the same at every point around the beacon in the case of no modulation. For a phase difference of less than 90 deg. the locus of the rotating vector would be elliptical and the pattern of field intensity would have two minima which become less sharply defined as the phase difference increases. Observations taken by means of a thermoelement at a distant receiving point, as the goniometer is rotated, can

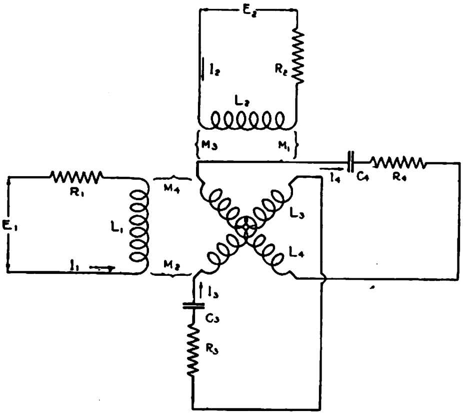


Fig. 14.—Circuit Relations in Goniometer—Antenna System.

thus be used as a ready check upon the equality of phase adjustment of the antenna currents.

With modulation impressed on the two antenna currents, no difficulty is encountered by reason of a phase displacement, provided each modulating frequency is entirely confined to its own antenna.

The matter of inequality of these phase angles, however, is important, as it introduces reaction back into the primary coil circuits. Referring to Fig. 14, L_1 and L_2 are the input goniometer coils, and L_3 and L_4 those connected to the antennas. As the two input circuits are identical with the same voltage impressed on each of them, we can put

$$M_1 = M_4 = M\cos\theta$$

$$M_2 = M_3 = M\sin\theta$$

$$R_1 + j\omega L_1 = R_2 + j\omega L_2 = Z$$

$$R_4 + j\left(\omega L_4 - \frac{1}{\omega C_4}\right) = Z_4$$

$$R_3 + j\left(\omega L_3 - \frac{1}{\omega C_3}\right) = Z_3$$

The following relations can be written

$$E_1 = I_1 Z + j\omega M I_4 \sin\theta + j\omega M I_3 \cos\theta \tag{9}$$

$$E_2 = I_2 Z + j\omega M I_4 \cos\theta + j\omega M I_3 \sin\theta \tag{10}$$

$$0 = I_3 Z_3 + j\omega M I_1 \cos\theta - j\omega M I_2 \sin\theta \tag{11}$$

$$0 = I_4 Z_4 + j\omega M I_1 \sin\theta + j\omega M I_2 \cos\theta \tag{12}$$

From (11) and (12) we get

$$I_3 = \frac{j\omega M}{Z_3}(I_2 \sin\theta - I_1 \cos\theta) \tag{13}$$

$$I_4 = \frac{j\omega M}{Z_4}(I_2 \cos\theta - I_1 \sin\theta) \tag{14}$$

If the impedances Z_3 and Z_4 are made equal, the antenna currents will be in phase. This condition is fulfilled when both circuits are in resonance, and the values of I_1 and I_2 become after placing values for I_3 and I_4 in (9) and (10)

$$I_1 = \frac{E_1}{Z + \dfrac{\omega^2 M^2}{Z_0}} \tag{15}$$

$$I_2 = \frac{E_2}{Z + \dfrac{\omega^2 M^2}{Z_0}} \tag{16}$$

where Z_0 is the antenna impedance, or in the case of resonance its resistance. Here the values of currents I_1 and I_2 are independent of the coil angle. It can be seen in this case that the current in each primary circuit is due only to the applied voltage of its own circuit.

If $E_1 = E_2 = E$ and Z_3 is not equal to Z_4 in which case the antenna currents are not in phase, the solutions for I_1 and I_2 are

$$I_1 = E\frac{\left(Z + \dfrac{\omega^2 M^2}{Z_4}\right) + \dfrac{\omega^2 M^2}{2}\left(\dfrac{1}{Z_3} - \dfrac{1}{Z_4}\right) + \dfrac{\omega^2 M^2}{\sqrt{2}}\left(\dfrac{1}{Z_3} - \dfrac{1}{Z_4}\right)\sin\left(2\theta - \dfrac{\pi}{4}\right)}{\left(Z + \dfrac{\omega^2 M^2}{Z_4}\right)\left(Z + \dfrac{\omega^2 M^2}{Z_3}\right)} \tag{17}$$

$$I_2 = E\frac{\left(Z + \dfrac{\omega^2 M^2}{Z_4}\right) + \dfrac{\omega^2 M^2}{2}\left(\dfrac{1}{Z_3} - \dfrac{1}{Z_4}\right) + \dfrac{\omega^2 M^2}{\sqrt{2}}\left(\dfrac{1}{Z_3} - \dfrac{1}{Z_4}\right)\sin\left(2\theta + \dfrac{\pi}{4}\right)}{\left(Z + \dfrac{\omega^2 M^2}{Z_4}\right)\left(Z + \dfrac{\omega^2 M^2}{Z_3}\right)} \tag{18}$$

and it can be seen that these currents vary with the angular position of the goniometer primary and secondary coils to each other showing that one primary circuit reacts on the other primary circuit by way of the antenna circuits.

If E_1 and E_2 are each modulated to one of the desired low frequencies, such as 65 and 86 cycles, the expressions for I_1 and I_2 become much more complicated and I_1 will contain a component due to voltage E_2, and I_2 likewise will have a component due to voltage E_1, for all values of angle θ except certain multiples of $\frac{\pi}{4}$. This situation is equivalent to that produced by coupling between the amplifier systems which introduce distortion in the output characteristic. The maintenance of resonance in both the antennas therefore constitutes an important operating condition.

RECEIVING EQUIPMENT

The only airplane antenna system that has been found effective heretofore is the weighted trailing wire. This, generally speaking, has marked directive properties[4] so that unless it lies in the plane containing the transmitting station, erroneous indica-

found to be very satisfactory from all standpoints. No apparent errors in the beacon courses could be noticed. With the trailing wire it was a very difficult feat to guide an airplane right to the beacon, and it became impossible when a side wind produced a slide slant to the antenna. With the vertical pole antenna it became very simple to fly directly towards the beacon. Furthermore since a horizontal electric field induces no current in the vertical pole antenna and as no vertical field exists directly above the beacon station, there is a cessation of signal at the moment when the airplane passes over it. The beacon can thus be located within 100 feet when the airplane is not over 1000 feet above it, a most valuable aid to landing in fog.

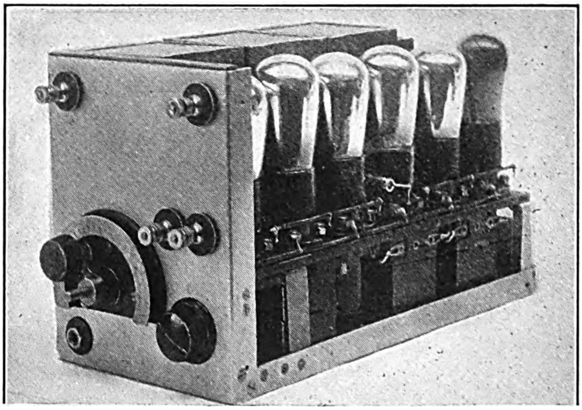


Fig. 16.—Special Receiving Set Developed for Aircraft Reception of Radio Beacon and Telephone Signals.

Receiving sets have been developed for this work which supply ample voltage for operating the visual indicator unit, with a radio-frequency input voltage of as low as 10 microvolts. These sets (see Fig. 16) are equipped with a uni-tuning control to cover the aircraft telephone and beacon frequency bands extending from 285 to 350 kc, allocated by the 1927 International Radiotelegraph Convention. The set has an audio-frequency amplifier which is efficient down to 50 cycles; it weighs from 12 to 16 lbs., depending upon the design. A small panel, mounted in the cockpit or on the instrument board, contains a volume control, battery voltmeter, filament switch, and jack for head telephones. The pilot adjusts the volume control to maintain a convenient amplitude of vibration of the reed indicator.

With sets of this sensitivity, full range of operation cannot be secured without careful engine ignition system shielding. Merely shielding the high and low tension wires is not sufficient. The spark plugs must also be screened. It seems that the advantages of the vertical pole antenna must be purchased at the price of installing spark plug shielding, which entirely eliminates the undesired interference. There is sufficient room on the modern radial type of airplane engine to accommodate spark plug covers, and a successful installation of this kind has been made at College Park. It is expected that suitable spark plugs and shielded wire harnesses to connect with them will be available on the market in the near future.

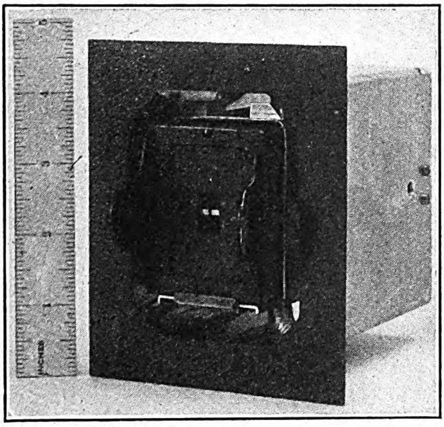


Fig. 17.—Reed Indicator Unit with Spring Holder

Figs. 17 and 18 show one type of reed indicator unit. This unit is designed to match the output impedance of the receiving set, which is between 4000 and 7000 ohms, and gives full scale deflections with from 4 to 10 volts across the terminals. The indicator reeds may be all steel, or may consist of a small steel armature with the principal vibrating portion of elinvar. The latter has the advantage that the reed tuning is practically independent of temperature because the modulus of elasticity of elinvar has a negligible temperature coefficient. When an all-steel reed is used, the effect of temperature in varying the tuning of the reed is such that damping must be provided to make it respond to frequencies over a range of plus or minus at least 0.4 per cent. The damping is introduced by so shaping the reed as to introduce an appropriate amount of surface moving against the air. With any type of reed some damping is necessary, in order to keep the reed operative in case small variations of frequency occur in the radio beacon transmitting apparatus.

The indicator unit is placed on the airplane instrument board in a shock-proof mounting. The indicator with mounting weighs only about a pound, rather less than a head telephone set which it replaces. This system thus ideally meets the requirement of minimum weight aboard airplanes. The type of indicator shown plugs conveniently into its mounting. The plug-in arrangement has several advantages. It facilitates changing indicators when the airplane is used on a route in which different modulation frequencies are used at the beacon. It is also a convenient practice to reverse the connections of the indicator whenever flying past the beacon by removing the indicator, turning it upside down, and plugging in again. Then the convention as to relative amplitudes of reed vibration remains unchanged. For this reason, the indicator has "From beacon" and "To beacon" engraved at opposite ends of its face.

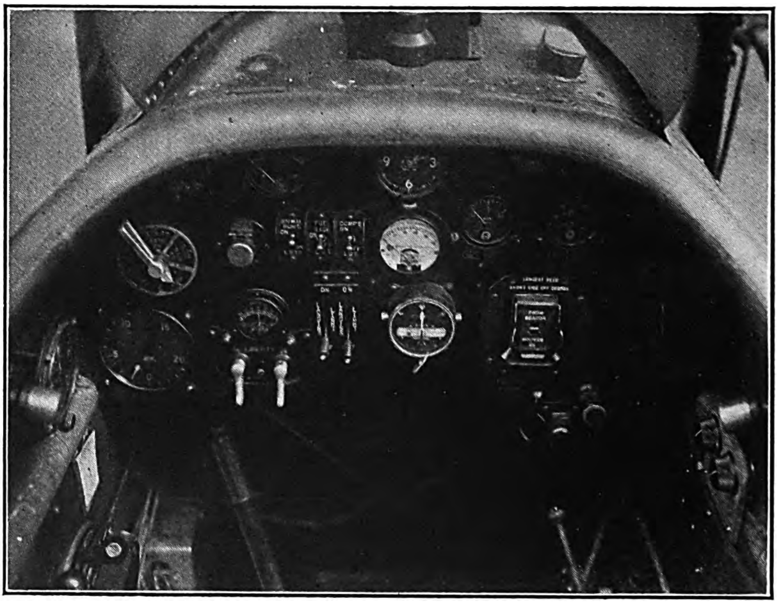


Fig. 18.—Reed Indicator Unit and Holder Mounted on Airplane Instrument Board.

DIRECTION VARIATIONS

The existence of errors at night in the apparent observed direction of radio stations has long been known.[8,9] Their existence in the case of the directive radio beacon of the type described was discovered by one of the authors[10] during test flights made in August, 1927.[11]

Investigations as to the causes of direction shifts have been under way for some time, particularly in England,[12] and experimental proof has been secured that these shifts are due to the horizontal component of the downcoming-reflected or refracted wave. Special experiments made upon our beacon stations have verified this.[11]

As it has been shown and experimentally verified, that the vertical pole type of airplane antenna is immune to horizontal electric fields when correctly installed, it is expected that with it, no night variations of magnitude will be observed.

PRACTICAL APPLICATIONS

By the use of the directive radio beacon, an airplane can be held on a straight course represented by a zone from 2 deg. to 5 deg. wide extending from the beacon station. The equipment described has a reliable day operating range of 100 miles. As can readily be seen, the controlled flight of the airplane along this straight course is independent of wind drift, so that the pilot does not need to take that into account.

However, there is no means whereby the pilot can judge his distance from the beacon, as a head or tailwind may alter the airplane's ground speed. Small marker beacons set at frequent intervals along the directive beacon course will provide this information. Tests have indicated that a 7.5-watt tube oscillator feeding a vertical antenna, both tuned to the same radio frequency on the directive beacon, will have a signal range of less than about

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[8] "Variation in Direction of Propagation of Long Electromagnetic Waves," A. H. Taylor, Scientific Paper, Bureau of Standards, No. 353, 1919.

[9] "Some Radio Direction-Finding Observations on Ship and Shore Transmitting Stations. Smith-Rose, R. L. *Journal* I.E.E. (London), **62**, pp. 701–711, August, 1924.

[10] Haraden Pratt.

[11] "Apparent Night Variations of Crossed Coil Radio Beacons," H. Pratt. PROC. I. R. E., **16**, p. 653; May, 1928.

[12] "The Cause and Elimination of Night Errors in Radio Direction-Finding." R. L. Smith-Rose and R. H. Barfield, *Journal* I. E. E. (London), **64**, pp. 831–843; August, 1926.

three miles. The present modulating frequencies used at the College Park station are 65 and 86 cycles. By supplying the marker beacon oscillator with a plate voltage secured from a commercial 60-cycle supply, a third reed unit on the airplane instrument board tuned to 60 cycles will respond to these markers, and if each marker emits a characteristic signal at an appropriate speed, it may be identified by the pilot. Here again, the marker will have above it a conical region of no signal, made possible by the vertical pole airplane antenna. This will make the marker's location capable of definite determination by the pilot.

The successful reception of a characteristic marker beacon signal with the reeds has demonstrated the feasibility of the sending of intelligence by telegraphic code over the system.

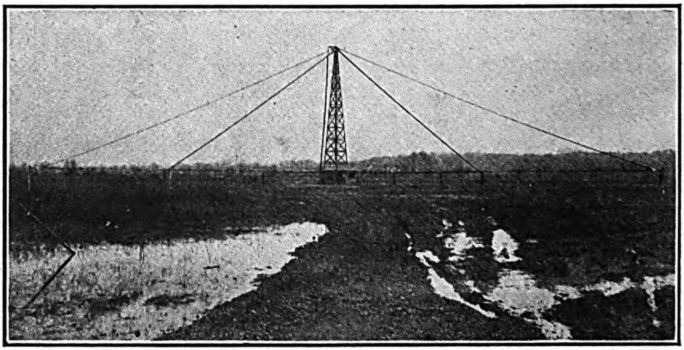

Fig. 19.—Directive Radio Beacon Station of Bureau of Standards, College Park, Md.

Weather information, for example, might be sent using a few simple code groups. Such information can be sent on the reeds provided for the beacon reception, or on one or more special reeds.

It has been found that if the two modulating frequencies of the directive beacon are too nearly alike, the low-frequency beats between them cause the reeds to flutter. This beat frequency must be sufficiently high so that the reed does not have opportunity to store up and dissipate energy in its oscillating state. With the system as now developed, a separation of 20 cycles between the reeds is necessary. This does not permit of more than two or, at the most, three groups of these frequencies to be used on the same carrier frequency in any one neighborhood.

This is not a serious matter as more than two or three groups in one locality will not be needed.

Since the marker beacon frequency of 60 cycles will not have a separation of 20 cycles from the 65-cycle modulation of the directive beacon, and may not have this separation from some other directive beacon frequency which may be in operation in the vicinity, some flutter will occur when very close to the marker. Tests have indicated that the degree of flutter in that case is not of a serious nature as the marker beacons use low power and have a short distance range.

The plugging of a pair of head telephones, or an aviation helmet equipped with telephone receivers, into the output circuit of the receiving set, makes possible the reception of radio telephone messages. The change to the telephone connection involves an adjustment of the unicontrol tuning element on the receiving set at present in use. At the present time the College Park and Bellefonte beacons are adjusted to a frequency of 290 kc and the ground telephone transmitting stations to the new International calling and distress aircraft frequency of 333 kc. Suitable stops on the receiving set tuning control lever adjusted for these frequency settings enable the pilot to make the change from one to the other with a minimum of effort.

With a ground station power of 1 kw, reliable telephone signals have been received on an airplane in flight for a distance of 100 miles. The important limiting factor is the engine ignition interference, as noise from this source that will not influence the reed indicator will create serious interference with the reception of speech. The practical distance range for speech is also somewhat reduced by the presence of a considerable general noise level on the airplane, which is not a factor in the operation of the reed indicator. Experience to date has indicated that a somewhat greater distance range may be expected with the visual indicator than for radio telephony, under similar conditions of power and receiving equipment.

The combination of directive and marker beacons with weather and other information broadcast to airplanes by radio telephony, properly organized, thus provides a complete set of radio aids for air navigation. They permit flying under conditions of no visibility, and should add materially to the safety and reliability of air transportation.

In concluding, the authors desire to acknowledge the parts played in these developments by their coworkers in the Bureau of Standards. Appreciation of valuable contributions is extended to F. W. Dunmore, particularly for ingenious laboratory developments and basic ideas in connection with the modulation type of beacon and the visual indicator; to Harry Diamond for the design and construction of the airplane receiving sets, and the research incident to the operation of the beacon and its modulating arrangements; and to Dr. E. Z. Stowell for contributions to the earlier circuit arrangements described and studies of the field intensity diagrams. Acknowledgments are also due to Captain Russell L. Meredith, airplane pilot, C. B. Hempel and R. M. Green for field installations and tests, F. G. Gardner, airplane mechanic, and L. L. Hughes, D. O. Lybrand, and R. R. Gessford, who manufactured much of the apparatus used.

AIRCRAFT RADIO INSTALLATIONS*

By

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Summary—This paper deals in a general manner with the technical aspects of aircraft radio design and installation, and illustrates the trend of development during recent years. Radio conditions and installations aboard dirigibles (lighter-than-air craft) are described and compared with the more severe conditions and requirements aboard airplanes (heavier-than-air craft). The bearing of these special conditions both on receiver and transmitter design is shown, resulting in highly specialized equipment for aircraft use, with several distinct means of power supply. Described and illustrated are also various expeditionary aircraft radio installations, emergency radio equipment, some special high-frequency aircraft radio apparatus, as well as some standard installations in naval planes. Advantages and disadvantages of various directional radio applications for aircraft navigation are shown. The paper concludes with a brief description of recently developed commercial radio equipment, and an enumeration of ground radio facilities being installed by the government; these are expected to have a far-reaching influence on safety and dependability of our rapidly developing commercial air transportation.

A. Introduction

AFTER a period of rapid growth, aviation today has reached a state where its successful commercial application is as much concerned with development of essential accessories as with improvements in the aircraft itself; thus one of the greatest needs today of commercial air transportation is the provision of simple and accurate means of navigation and of safety in landing under unfavorable flying conditions with poor visibility. The importance of radio communication to aircraft is apparent; its value has been proved not only in many years of application in our military services but also by foreign everyday commercial use, as well as on a number of more or less spectacular long-distance flights. If existing radio possibilities have not been realized or taken advantage of on some otherwise well-prepared flights, this has been due in many cases to lack of information on aircraft radio by the flight organiztion; in some cases, the operating or installation personnel has not

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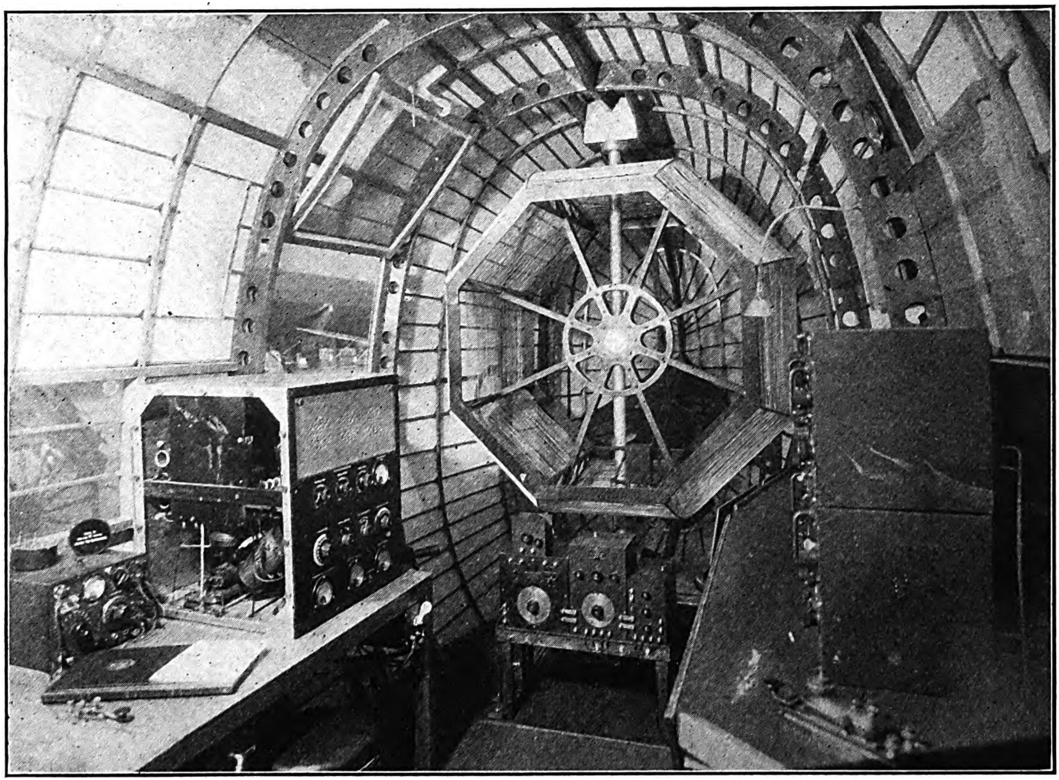

Fig. 1—Looking Aft in Radio Room, USS *Shenandoah*.

been sufficiently competent, or the equipment has been unsuitable. It is surprising how little is known about the special problems and developments of aircraft radio even among professional radio engineers. It is impossible within the scope of this paper to more than touch upon the various requirements and conditions encountered in this large special field, but it is hoped that this paper will clear up some past misapprehensions and result in a better understanding of this important application of radio.

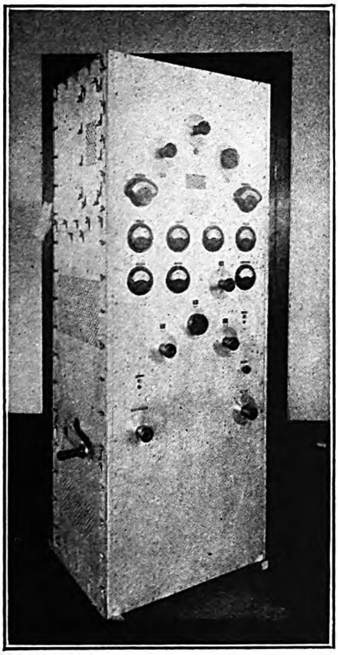

Fig. 2—Front of 2 kw. Transmitter Built for USS *Shenandoah.*

B. Radio Equipment for Dirigibles

Conditions met with on aircraft demand that most radio equipment carried be of highly specialized design, both electrically and mechanically, in order to meet the severe requirements imposed. Dirigibles, or airships, so-called lighter-than-air craft, beyond imposing space and weight restrictions on the equipment are relatively free from the detrimental conditions found on heavier-than-air planes; in fact, the height of the radiating system above the energy-absorbing ground together with the effective low-resistance counterpoise offered by the all-metal framework give the large rigid dirigibles advantages not found on ground or shipboard installations and make possible communication over great distances with radio equipment of necessarily restricted power and weight. Fig. 1 shows a typical installation in a large airship, the rear portion of the radio room in the late USS *Shenandoah*. In order to provide a relatively unshielded location for the radio compass shown in the center of the picture, the radio room was of wood and fabric construction and in this respect differed from the metal framework construction of the remainder of the ship including the control room ahead. Being a part of the forward "car" or gon-

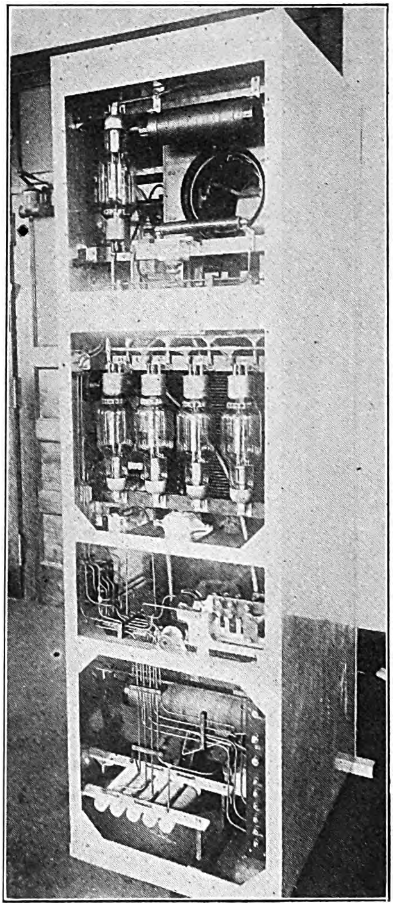

Fig. 3—Right Side of 2 kw. *Shenandoah* Transmitter.

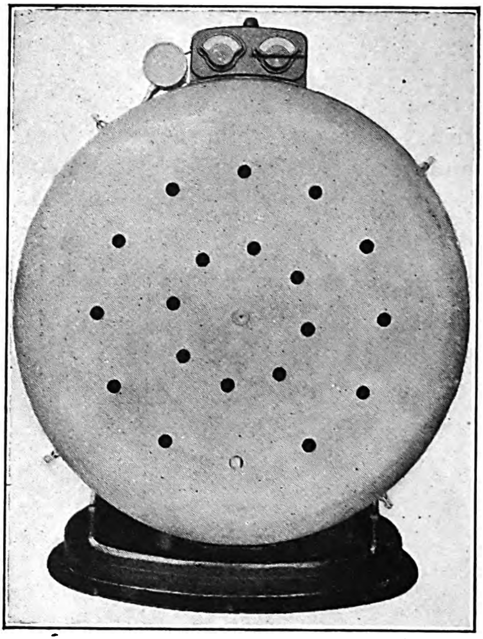

Fig. 5—Front of Gasoline Driven Generator for USS *Shenandoah.*

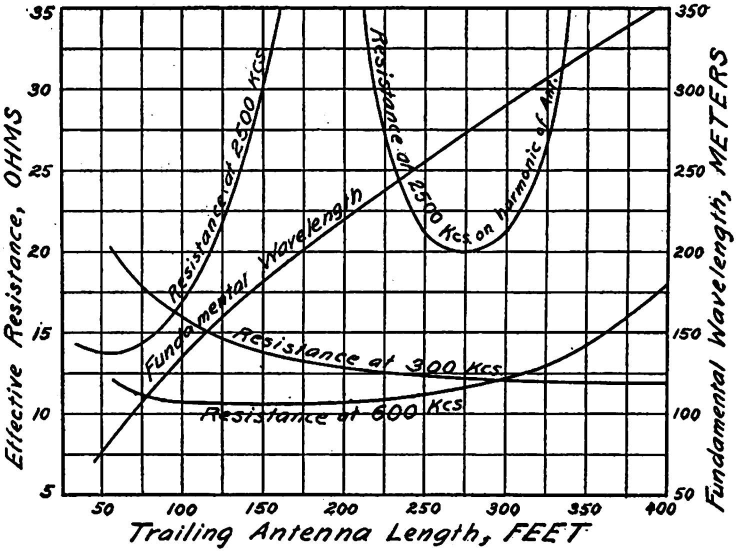

Fig. 6—Effective Resistance and Natural Wavelength Curves for Single Wire Trailing Antennas.

ated intermittently. Undue noise was avoided by enclosing the power supply equipment and batteries in a small separate compartment between the radio room and the control room.

When plans were made in 1923 to send the *Shenandoah* on an Arctic exploration flight, the preparations included design and construction by the Naval Research Laboratory of a long range radio transmitter, probably the largest and most powerful set ever built for aircraft. This transmitter is shown in Figs. 2 and 3. Notice the large antenna reel built into the lower part to save space. ACW tone telegraphy with about two kilowatts antenna input was provided by eight 250-watt power amplifiers excited by two quarter-kilowatt master oscillators, at frequencies between 200 and 500 kc. The power supply for this transmitter is shown in Figs. 4 and 5; it consisted of a specially developed 15 H.P. gasoline-electric unit built by the Kinney Mfg. Co. of Boston, and delivered 5 K.V.A. at 220 volts, 660 cycles alternating current, as well as 2 kw. at 30 volts direct current for field excitation and battery charging. This unit was supported on a pneumatic cushion to reduce vibration

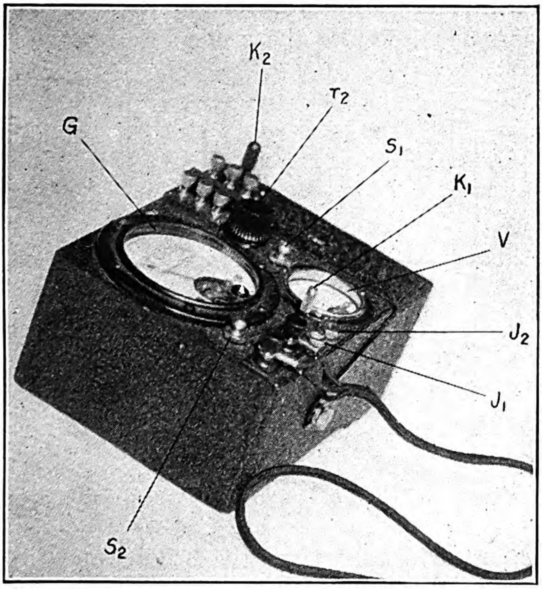

Fig. 7—Visual Signal Intensity Meter.

and noise. The weight of the transmitter unit shown was about 425 lbs. inclusive of tubes and antenna; the power supply unit weighed about the same. The normal telegraph range of this equipment was expected to be about 1500 miles in daylight. Because of abandonment of the Polar flight plans this powerful set was never installed in the ship, and subsequent developments in high-frequency equipment have made airship transmitters of this size and weight unnecessary for long distance communication.

The west coast cruise of the *Shenandoah* in the fall of 1924 gave an opportunity to test the possibilities of high-frequency aircraft equipment; despite the crude receivers of those days, two-way communication between the dirigible and the Naval Research Laboratory was accomplished night after night on the entire trip, the Laboratory employing a 54-meter transmitter with about 500 watts in the antenna, and the *Shenandoah* transmitting on 90 meters with 50 watts output. The USS *Canopus* reported instances of good night reception of the *Shenandoah's* signals at Guam while the airship was flying in the vicinity of Seattle, 5000 miles away.

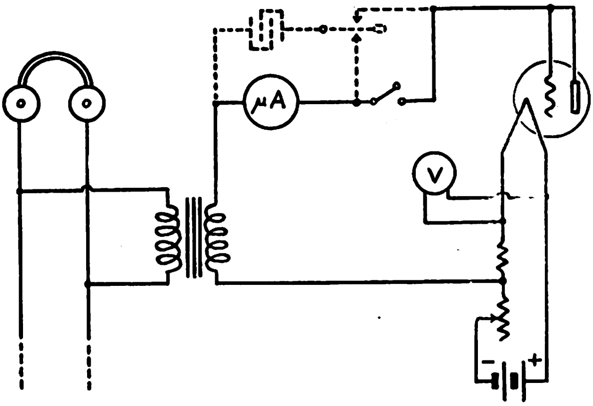

Fig. 8—Schematic Circuit Diagram of Visual Intensity Meter

Our only present rigid dirigible, the USS *Los Angeles*, is equipped with a German Telefunken intermediate-frequency transmitter of 200 watts antenna input, and a high-frequency 50-watt crystal-controlled transmitter operating on 3475 and 8012 kc.; either of these transmitters has a daylight range of approximately 500 miles. In addition, the *Los Angeles* is equipped with a rotating coil radio compass, a German plug-in coil universal receiver, and a high-frequency receiver. Radio operating conditions on these dirigibles have been found very advantageous. In addition to the trailing-wire antenna employed, a short fixed antenna is provided which is especially valuable for communication with the ground crew in landing operations.

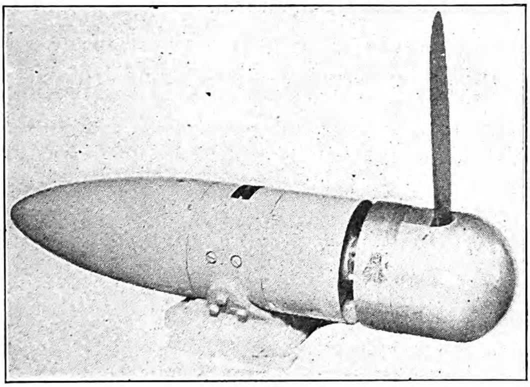

Fig. 9—R.C.A. Wind-Driven Radio Generator with Deslauriers Self-Regulating Propeller.

C. General Flight Conditions Affecting Radio

As previously mentioned, in view of the favorable radiation conditions in flight good results are usually attained by use of a single trailing wire antenna, against the metal structure and bonding of the plane as counterpoise; this gives a good effective height with relatively little absorption loss. An idea of the

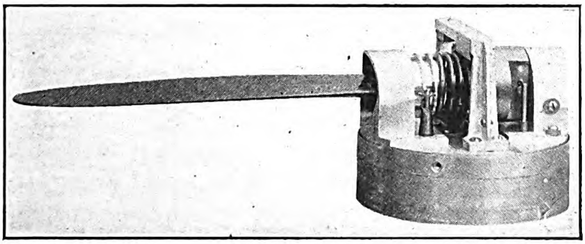

Fig. 10—Mechanism of Deslauriers Self-Regulating Propeller.

effective resistance and the fundamental wavelength of such a trailing antenna is obtained by reference to Fig. 6, which is based upon average values observed on planes of moderately large size. As indicated by the 2500 kc. resistance curve, it is practicable and sometimes desirable at the higher frequencies to operate on harmonics of the antenna system; antenna lengths employed for intermediate frequencies on airplanes are sometimes as much as 350 feet or even more, but modern tendency with efficient transmitters is rather to avoid great lengths, for reasons of safety and convenience. Stranded phosphor bronze or copperclad steel wire is generally employed, carrying a total weight of from two to five pounds at the end. To reduce the possibility of accidental loss and to make possible ready renewal of an antenna in flight, tubular antenna weights together with a large diameter fairlead to make them completely retractable are often used.

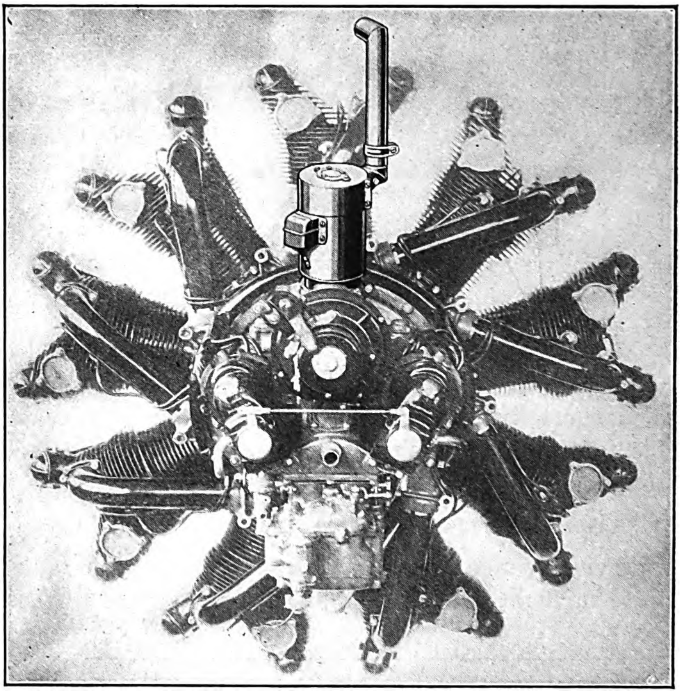

Fig. 11—Leece-Neville Generator Geared to Pratt and Whitney "Wasp" Engine.

Space and weight restrictions imposed on all aircraft radio equipment are usually severe and can be met only by skilful design; all equipment should be readily accessible for inspection and maintenance and must be constructed to withstand continued vibration and landing shocks without breakage. Equipment is generally supported on cushions of sponge rubber or on spring suspensions, or is hung in place by a suspension of rubber exerciser cord, also known as "Bungee" cord. Too resilient a suspension, however, is unsuitable as it may allow the equipment to bounce around and suffer severe shocks from striking an adjacent object during a bumpy landing.

A precaution which is necessary for efficient radio operation and which is also desirable for general safety reasons is the proper bonding of aircraft. By this is meant a thorough electrical interconnection of all metallic parts of the plane or airship, with permissible exception only of such isolated parts as actually are well-insulated from the main metal structure. Bonding is effected by means of copper straps or pigtails, with particular attention to clean and durable connections. The purpose of bonding is three-fold: First, it reduces fire hazard by preventing sparking between adjacent metallic parts. In the absence of bonding such sparks may occur as the result of charges of atmospheric electricity or from voltages induced by radio transmission. Obviously, especial care must be exercised to bond all fuel tanks and feed pipes with adjacent metal work; on hydrogen-filled dirigibles, bonding of all metal parts in the vicinity of the gas cells or envelope is a safety precaution of the greatest importance. A second advantage of bonding is the resulting increase of effective counterpoise area for radio transmission, and reduction of the radio-frequency resistance.

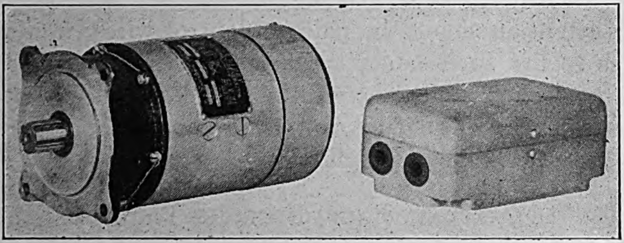

Fig. 12—Eclipse Engine Mounting Generator with Voltage Regulator.

Thirdly, absence of bonding may cause many electrical disturbances in radio reception, resulting from intermittent slapping or rubbing together of separate conductors in the field of the antenna-counterpoise system. To avoid such noises particular attention should be paid to control wires; it is modern practice to run such cables through sheaves and guides which are either well-grounded or well-insulated and to cover one or both cables with insulating sleeves at points where they cross and may slap together in flight.

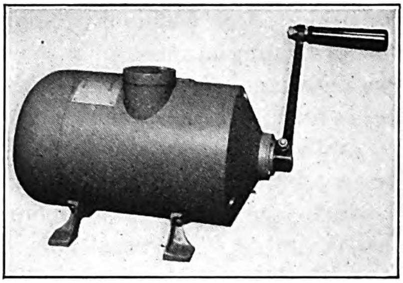

Fig. 13—Evershed's Hand-Driven Radio Generator, with Automatic Contactor.

D. Aircraft Radio Receiver Problems

There are many conditions adversely affecting radio reception in flight; these are only recently being overcome in a satisfactory degree, by specialized design and installation precautions. One of the worst obstacles is the great noise produced by the motor, usually accompanied by noise from the propeller and whistling of exposed wires. These noises, which are naturally more pronounced in open planes, can be partly

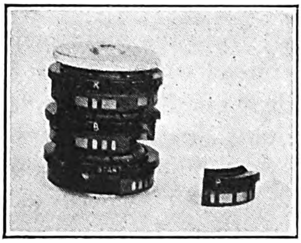

Fig. 14—Code Disk Arrangement in Frost's Automatic Contactor.

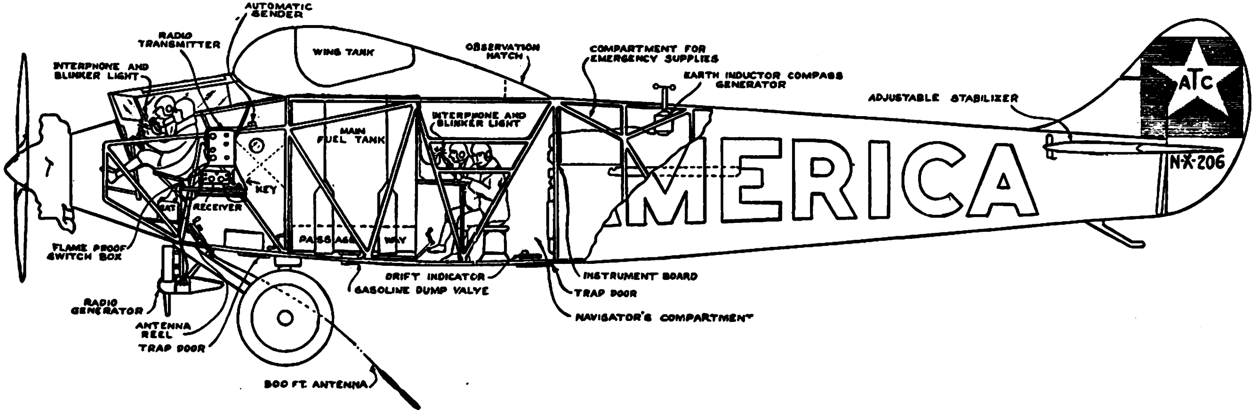

Fig. 16—Interior Arrangement of Byrd's Airplane *America*.

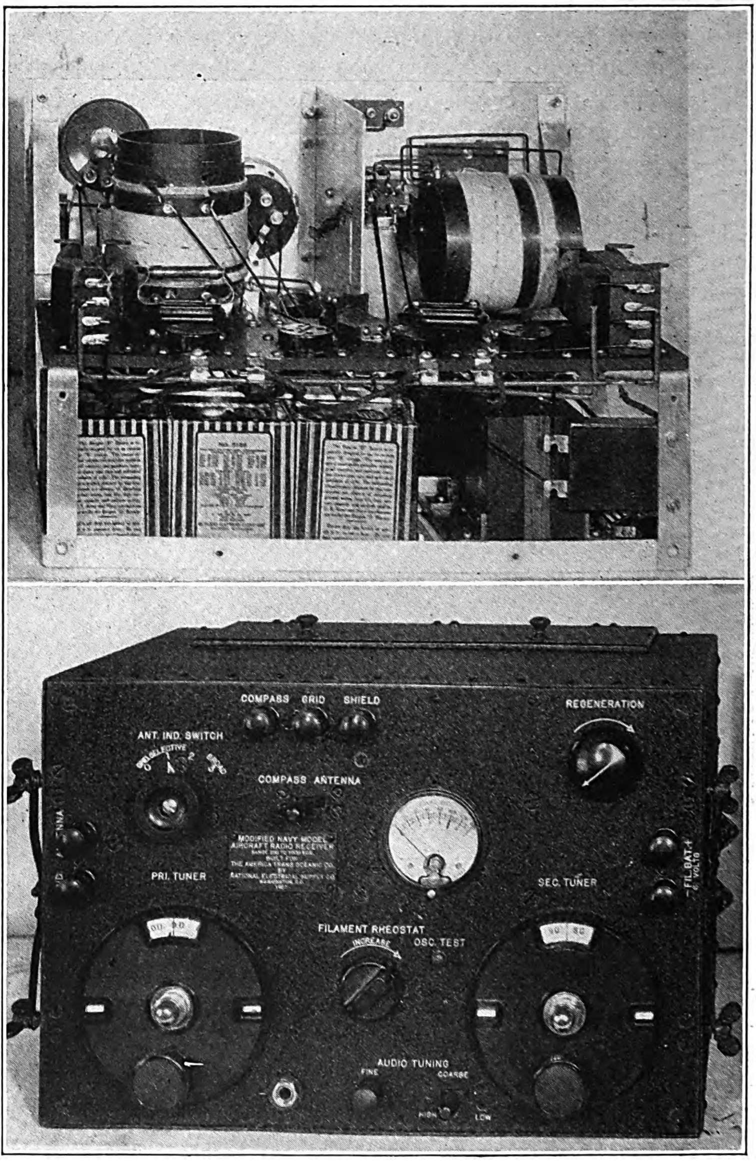

Fig. 18—Radio Receiver Used on Byrd's Transatlantic Flight.

considerably minimized by special anti-noise microphones which are sensitive to voice but exclude or balance out most extraneous sounds.

The power supply for the transmitter radio generator may be derived directly from the airplane motor by gearing, or indirectly by fan drive in the airstream, or from a storage battery through a suitable dynamotor. The latter method will readily furnish full power on the ground or water as well as in flight, but permits operation in flight for a limited time only, unless the battery is kept charged by another generator, such as has been done on many Army installations. The wind-driven generator with self-regulating fan gives the most flexible installation and will furnish power as long as the plane is in flight; if suitably installed in the propeller slipstream, the wind-driven

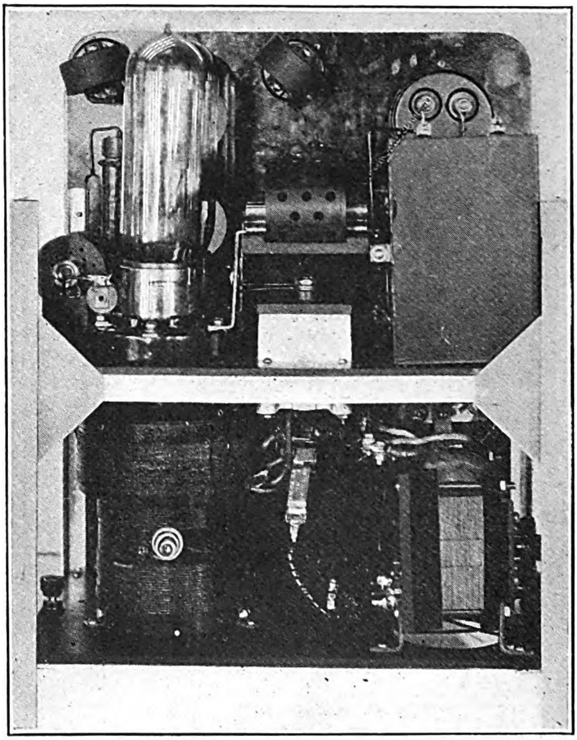

Fig. 20—Rear View of *America* Transmitter.

generator can also be made to turn up on the ground, for testing or for emergency communication. Fig. 9 shows a streamlined radio generator of several hundred watts output, supplying filament and plate voltages, and driven by a self-regulating air propeller. In Fig. 10 is shown the centrifugal governing mechanism which regulates the pitch of the counterbalanced single blade to keep the speed of rotation constant. Fig. 11 shows a 12-volt, 15-ampere generator mounted on an air-cooled engine; notice the ventilating stack for cooling the generator. Fig. 12 shows a somewhat similar generator with its mounting flange, and the box containing a vibrating voltage regulator as well as a battery cut-out. Such generators, for radio purposes, are provided with separate commutators for plate and filament voltages and are now manufactured by several concerns.

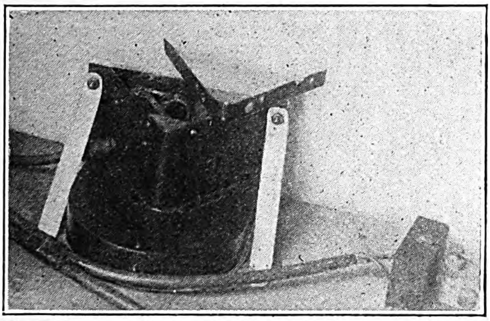

Fig. 21—Wind-Driven Call Disk Installed in Airplane *America.*

A type of current supply especially suitable for emergency use is the hand-driven radio generator shown in Fig. 13. This device, of British make, furnishes up to 50 watts of combined electrical energy for plate and filament supply; it has been found that one man is not able to exceed this power for any length of time, and in view of this fact, such hand-driven generators lend themselves especially to high-frequency emergency transmitters, which may despite their low power reach out over great distances. Fig. 14 shows an automatic contactor incorporated in some models of the hand generator, enabling automatic transmission of any chosen letter combination as the generator is cranked.

In another part of this paper is described a high-frequency transmitter which effectively uses dry batteries as power source.

F. COMPARISON OF RADIOTELEPHONY WITH RADIOTELEGRAPHY

Because of its convenience and in view of personnel limitations it is likely that radiotelephony will continue to be employed extensively for many classes of aircraft communication, up to distances of 100 miles or more. Where accuracy of communication, simplicity of equipment, or distance range without excessive power are requirements, however, radiotelegraphy offers advantages, provided suitable operating personnel is available. In Fig. 15 are shown curves, based on average American and foreign practice, which compare average intermediate frequency communication range with antenna power and total installation weight, for radiotelegraph and radiotelephone equipment. Practice has shown that for a given antenna power and under average conditions, well-modulated radiotelephony will

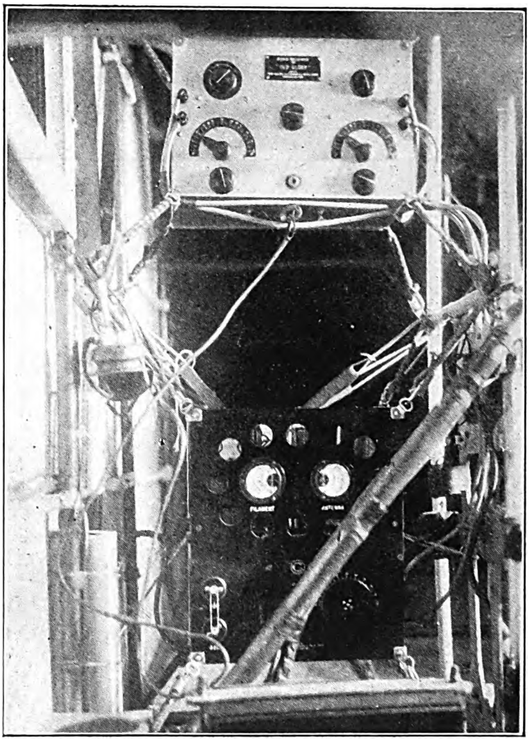

Fig. 22—Radio Installation by Allen D. Cardwell Co., in Airplane *Old Glory*, Front View.

construction, but the tubes and circuits may be safely loaded to a greater extent in view of the intermittent keying in place of a continuous carrier and modulation current.

G. SOME EXPEDITIONARY INSTALLATIONS

Long-range telegraphic equipment in its simplest form and of lightest weight is required for many expeditionary and long-distance flights. Where such flights are over long stretches of water an intermediate frequency without skip-distance and near the ship calling wave has obvious advantages. A typical installation of this sort was carried on Commander Byrd's flight across the Atlantic, installed in his airplane *America* as shown in Fig. 16. The equipment was patterned after installations which have proved their value on Naval scouting planes, but was modified in several respects for the sake of lightness, simplicity, and dependability, with the results that it functioned without failure during the entire forty-one hours that the plane was in the air. The wind-driven radio generator shown in Fig. 17 was a standard Navy type delivering 500 watts at 200 volts, 400 cycles, with the fan adjusted to 4000 r.p.m. Its location under the fuselage and near the landing wheels was prompted by necessity but exposed the fan to possible danger from flying stones and mud particles during take-off, and from swinging antenna weights while reeling in. Fig. 18 shows the receiver with self-contained plate battery which employed four tubes, namely, one stage of tuned neutralized r.f. amplification followed by a regenerative detector and two stages of audio amplification; a tuning range from 200 to 800 kc. was provided. Audio tuning of variable pitch could be switched across the

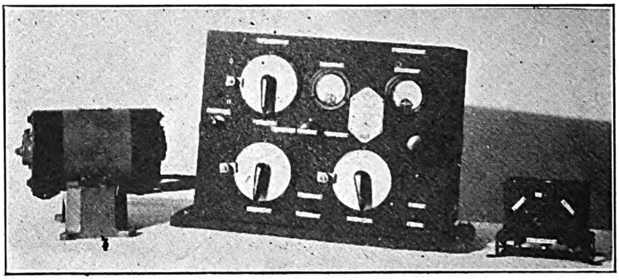

Fig. 24—Combined 600- and 45-Meter R.C.A. Transmitter Built for Airplane *American Legion.*

detector output if desired, and microphonic tube noises were reduced by acoustic damping with sponge rubber, sound shielding with lead foil, and damped flexible tube supports. Fig. 19 shows the front panel of the transmitter, which employed two 50-watt tubes, type UX-211 and delivered approximately 150 watts to the antenna. A full-wave self-rectifying circuit was employed with a transformer giving a plate voltage of about 1500 r.m.s. on either side of center tap. An antenna variometer enabled tuning to either a 690-meter working wave or a 600-meter calling and emergency wave, resonance to either being indicated by one of two small fixed wavemeter circuits with glow lamps, located in the upper portion of the panel. Fig. 20 shows the rear of this transmitter with the wavemeter circuits. In Fig. 21 is seen an innovation which proved of great value on this flight, namely, a wind-driven automatic code disk which continuously repeated the *America's* call letters WTW at times when the operator was engaged with non-radio duties; by this means stations and ships within range were enabled to keep track of and take bearings on the *America's* signals at all times and were kept on the alert for any messages and communications.

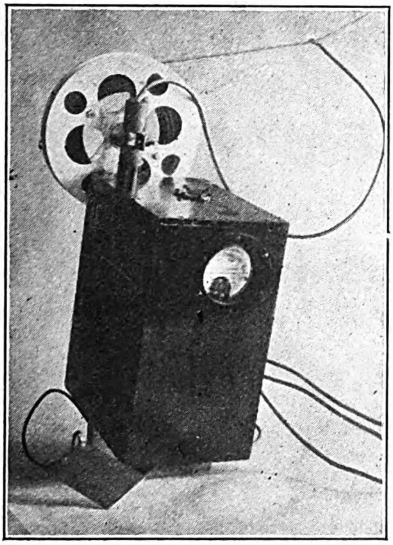

Fig. 25—Waterproof Self-Contained Emergency Spark Coil Transmitter.

Fig. 22 shows an installation somewhat similar to Byrd's in the airplane *Old Glory*, which in the summer of 1927 met disaster on an attempted New York to Rome flight. The installation is more fully shown in Fig. 23; the transmitting key is atop the flame-proof switch box, and under it is the antenna reel with fairlead and tubular antenna weights.

A totally different transmitter suitable for long-distance flights is shown in Fig. 24; this set, developed by the Westinghouse Company for the Radio Corporation gave excellent results on its test flights in Commander Davis' and Lieut. Wooster's ill-fated *American Legion*. The transmitter is operated from a storage battery and dynamotor and provides for approximately fifty watts CW output on either 45 meters or 600 meters; the 45-meter wave is crystal-controlled, while in the 600-meter position the transmitter is self-oscillating. A dependable transmitter of this type in the hands of a competent operator will give effective long-distance communication on high frequency, while providing for contact with commercial ship and shore stations on the 600-meter wave. Another advantage of this wavelength combination is the possibility of

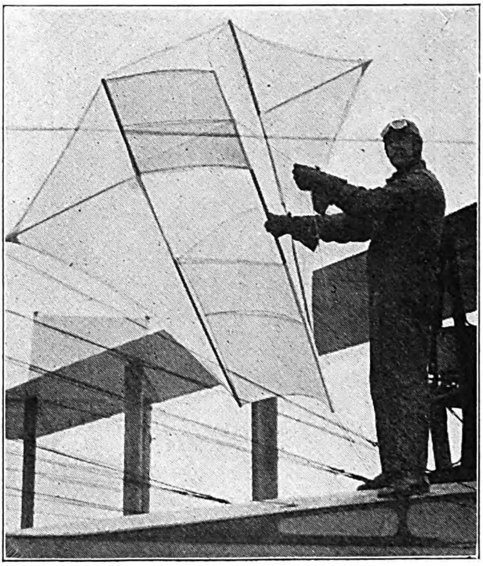

Fig. 26—Navy Emergency Radio Kite.

bridging with the 600-meter transmission any fading or skip zone likely to arise at moderate distances with the short wave.

H. EMERGENCY RADIO EQUIPMENT

For possible use in case of a forced landing, a completely waterproof and self-contained emergency transmitter was devised for Byrd's *America,* and so constructed that it could be operated from one of the inflated life rafts if required; such

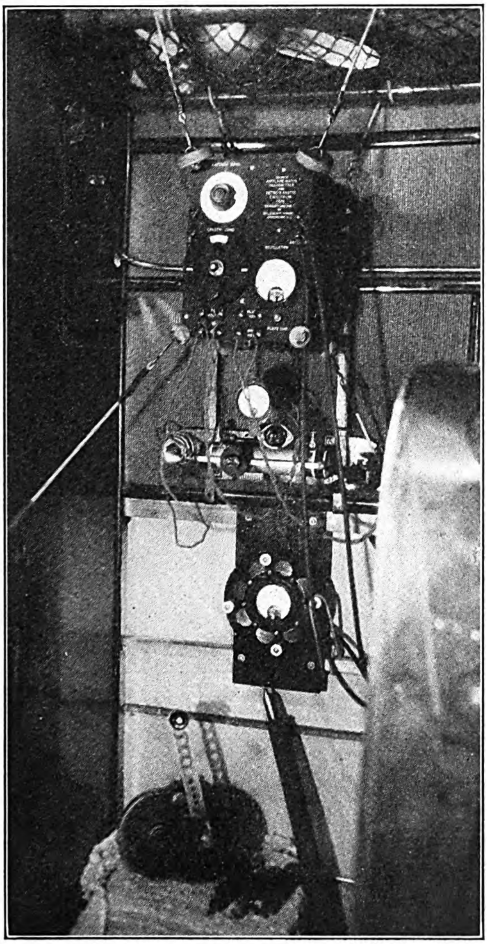

Fig. 27—High-Frequency Radio Transmitter Installation in Fokker Monoplane, Wilkins-Detroit Arctic Expedition, 1926.

transmitters were also carried by several other transatlantic flight contestants. This set, shown in Fig. 25, employed a spark coil operated from internally contained flashlight batteries which sufficed for several hours' telegraph operation; projecting waterproof leads made possible external connection to any other available battery. The small ground-plate shown was to be dropped into the water, while a 300-foot length of wire held aloft by a kite served as antenna. A radiation meter mounted behind a waterproof window indicated antenna current and resonance; tight inductive coupling with a fixed primary circuit tuned to 600 meters was employed. In tests, fair signals were still obtained from this transmitter at 25 miles distance. A standard Navy kite for carrying aloft an emergency antenna is shown in Fig. 26; such kites are provided in two sizes of 6 and 7½ feet height, for strong and light winds, respectively.

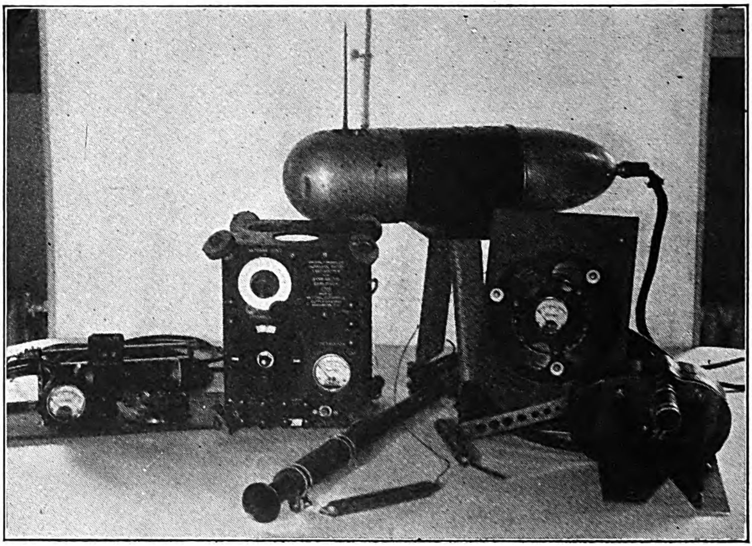

Fig. 28—Component Parts of Radio Installation Used by Byrd on Flight over North Pole.

Naval flying boats on the water employ for radio communication either a kite antenna or a fixed antenna supported by the wings and the tail structure. Power for the main transmitter may be obtained from a wind-driven generator in the slip stream of an operating propeller; hand-driven generators in conjunction with high-frequency transmitters have also been employed

heard as far as Unity, Saskatchewan, a distance of 1800 miles, and were reported also by a number of other stations beyond a thousand mile radius; this wave, however, showed a decided skip zone inside of 500 miles. A 40-meter wave, employed alternately with the higher frequency, showed less than 800 miles range but had no skip zone, although at close range fading was often pronounced and reception especially poor when the plane flew at low altitude. Reception of high frequencies in flight was found to be greatly hampered by ignition disturbance as well as vibration detuning and microphonic noises, but has since been accomplished effectively with suitably installed receivers of improved design.

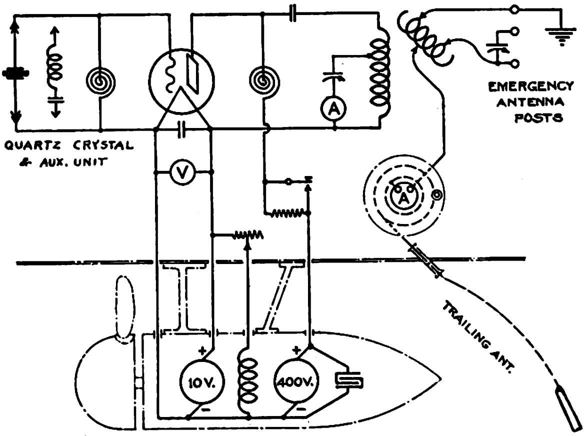

Fig. 30—Schematic Diagram of Crystal-Controlled High-Frequency Aircraft Transmitter.

When Captain George H. Wilkins conferred with the Navy in December, 1925, regarding suitable airplane equipment for his proposed North Pole flight to be attempted the following spring, the value of high frequencies for his purpose became apparent, and at his request a suitable airplane transmitter of very low weight was designed and constructed by the writer. This set, installed in Wilkins' single-motored Fokker plane, is shown in Fig. 27; an antenna input between 10 and 20 watts was obtained from a 50-watt Western Electric tube, with 400 volts on the plate, at crystal frequencies of approximately 5000 and 7000 kc. At Commander Byrd's request identical apparatus was constructed for his North Pole flight; this is shown in Fig. 28. Notice at the right the improvised hand gearing which was carried in the plane to replace the wind propeller in case of a forced landing. Fortunately Byrd did not need to employ this gearing on his flight; the few messages sent back to his base ship at Spitzbergen during the Polar flight

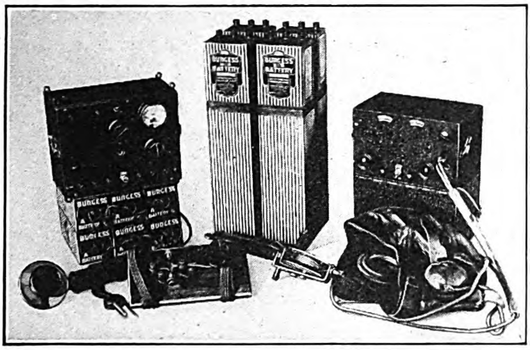

Fig. 31—Burgess Experimental Aircraft Set.

were received with good strength, but badly garbled by continuous spark interference from a schooner anchored nearby, carrying a flock of newspaper correspondents. More use was made of his radio equipment by Wilkins, who regularly communicated

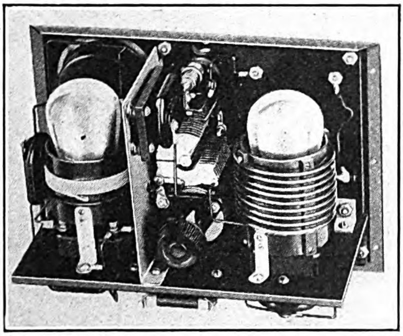

Fig. 32—Rear View of Interior, Burgess Airplane Transmitter.

over the 500-mile distance fron Point Barrow to Fairbanks, Alaska, even when turning the emergency gear from the ground. After two seasons' flight use, Wilkins in 1927 was finally forced to abandon this set on the polar ice with his damaged plane; after notifying the world of his forced landing, he and his pilot

Eielson had a month's struggle back to the mainland. He left behind with the set an Evershed hand generator which he had employed with good success. Fig. 29 shows the back of the Byrd-Wilkins set, with spare crystal and emergency self-oscillation unit, as well as emergency adapter and $7\frac{1}{2}$-watt tube for easier cranking in hand operation. The schematic diagram of the set is shown in Fig. 30; for ground operation, in place of the quarter or three-quarter wave trailing wire a simple fixed antenna could be improvised and tuned by means of

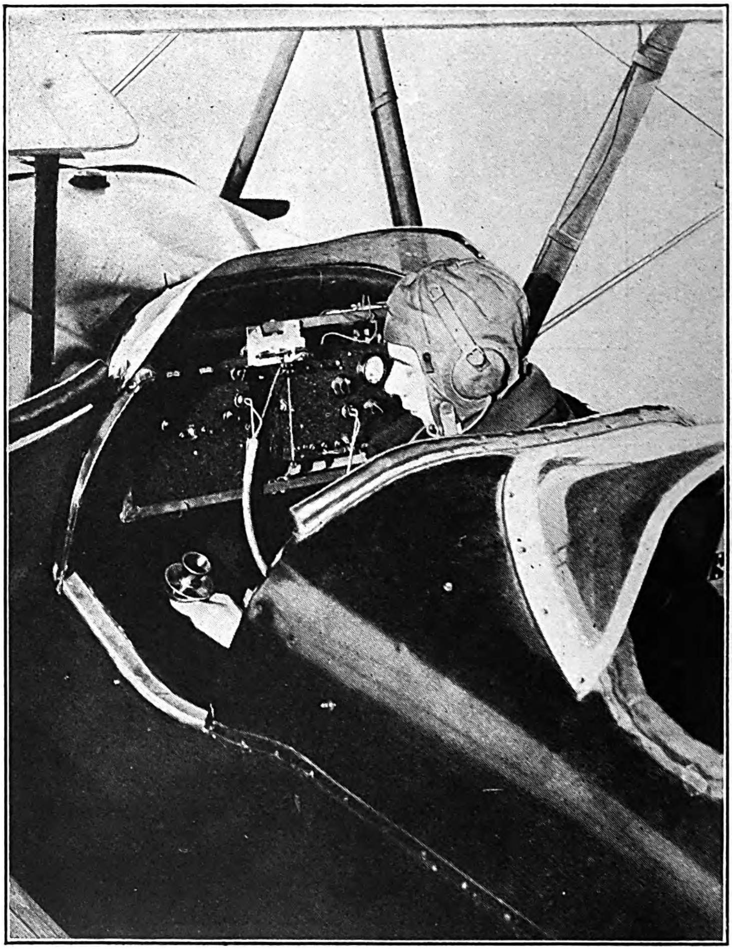

Fig. 33—Burgess Radio Equipment Installed in Travel Airplane.

a variable condenser connected in series or parallel with the antenna and counterpoise.

In view of excellent results obtained in point-to-point communication by Wilkins and others with portable high-frequency, battery-operated transmitting equipment of very low power, the Burgess Battery Company undertook a series of flight tests which resulted in the development of some extremely compact experimental aircraft radio equipment of interesting performance, shown in Fig. 31. Fig. 32 is an inside view of the little transmitter which employs one 201-A tube, shown at the left, as master oscillator, and a similar tube as balanced power amplifier. With 350 volts on the amplifier and 180 volts on the master oscillator plate, an antenna input of about 4 watts is reported; the master oscillator may be speech modulated by means of an absorption loop shown just beneath the white inductance winding. Installed in a Travel Air plane as shown in Fig. 33, experimental two-way communication with amateurs was accomplished on 79 meters up to 500 miles, and on 40 meters the plane's signals were reported up to a distance of 725 miles. Further contemplated tests with equipment of this

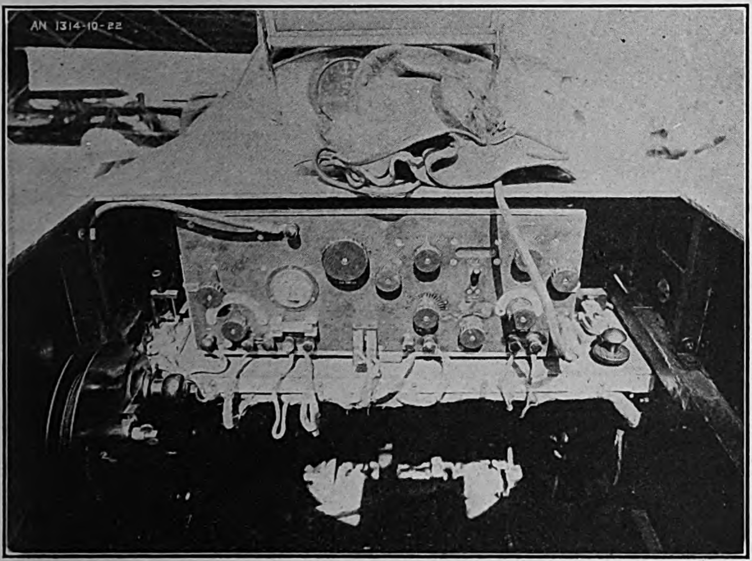

Fig. 34—Installation of Navy Type SE-1375 Transmitter-Receiver in Two-Seater Observation Plane.

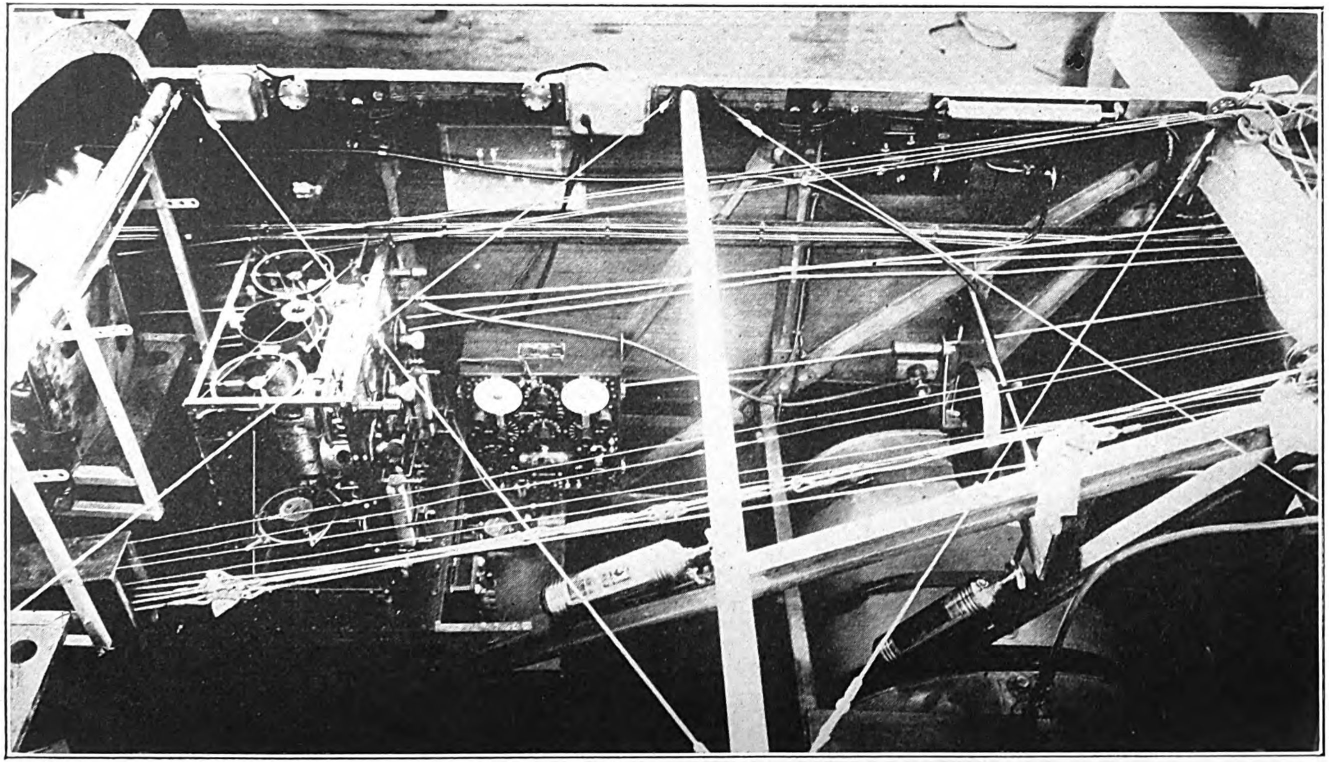

Fig. 35—War-Time Installation in Navy Flying Boat.

type are looked forward to with great interest to determine its possibilities in certain commercial applications. There is no question that for expeditionary purposes and uses requiring extreme portability, high frequencies offer great possibilities which in aircraft communication so far have only just begun to be realized. When the *Spirit of Dallas* went into a tail spin and disappeared into the Pacific Ocean last year, her SOS, transmitted on a 33-meter wave with a simple 50-watt set, was heard in New York City, and beyond as far as Italy. Extremely interesting results in high-frequency airplane experiments have also been obtained in Germany.[1]

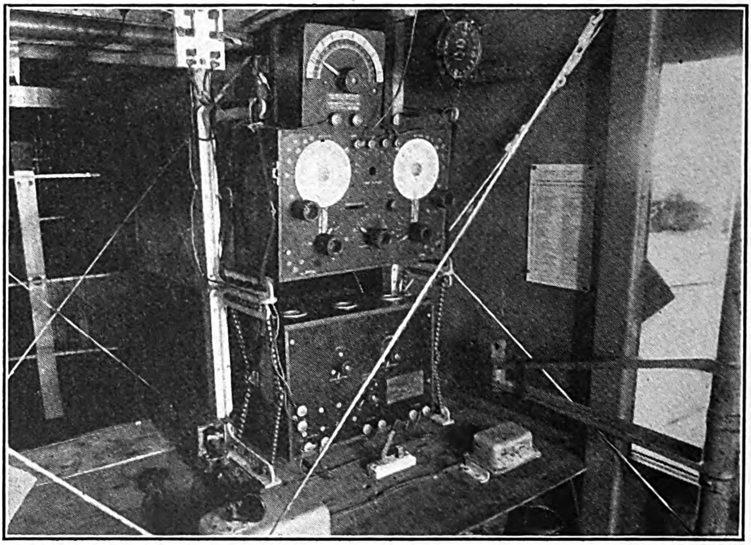

Fig. 37—Radio Compass Equipment in H-16 Type Flying Boat.

J. NAVAL AIRPLANE RADIO INSTALLATIONS

The early developments and extensive applications of Aircraft Radio in our Naval service are described in articles by T. Johnson, Jr.[2] Under the pressure of the war there was rapid and spectacular development; subsequently, there have been a no less important refinement, steady growth of application, and experimental research, which have resulted not only in in-

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[1] Plendl, H., "Die Anwendung Kurzer Wellen im Verkehr mit Flugzeugen." *Zeitsch. fuer Tech. Physik,* pp. 277-282, No. 7, 1927.

[2] Proc. I.R.E., 20, 1, 2; February and April, 1920.

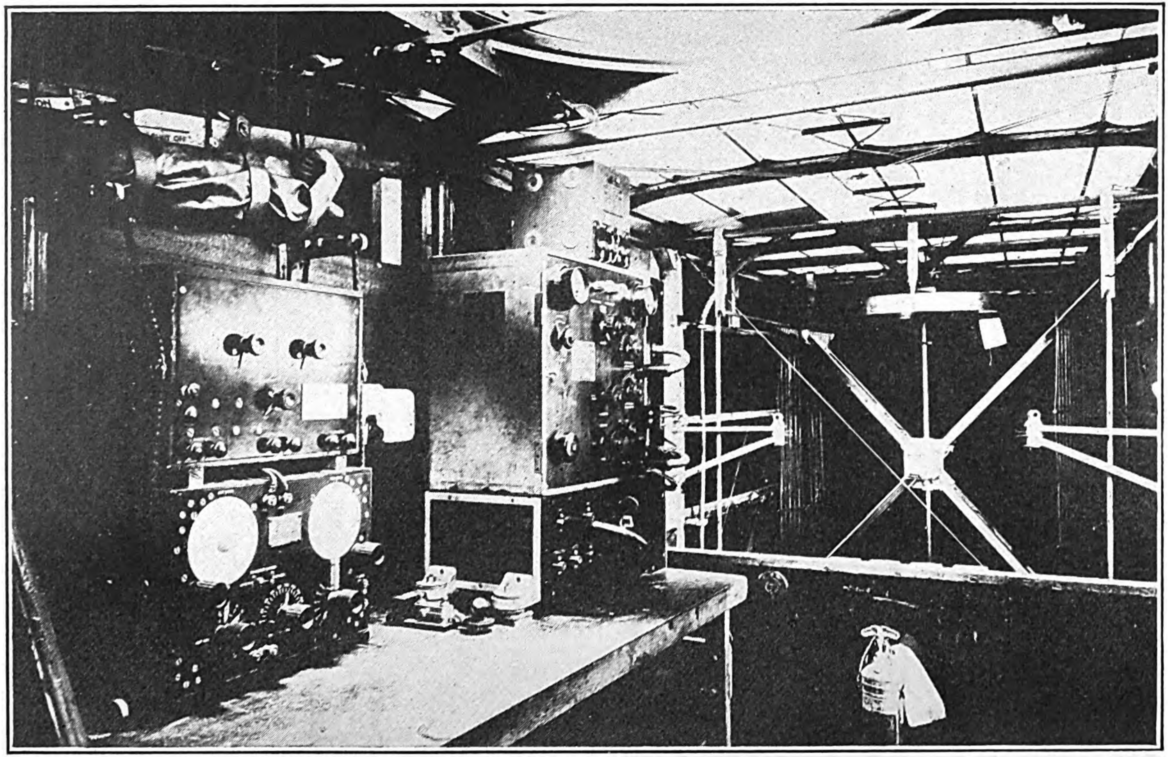

Fig. 39—Radio Equipment in PN-7 Flying Boat.

below. A 500-watt spark transmitter enclosed in a stream lined case with its generator was mounted outside, above the hull. The half passed directly over the masts of the ship, despite very poor conditions of visibility. Fig. 37 shows the radio compass equipment; note especially the shock proof receiver suspension on rubber cord. In Fig. 38 is shown also the combined telephone and telegraph 50-watt transmitter type SE-1390. Note also the pulley arrangement for rotating the compass loop.

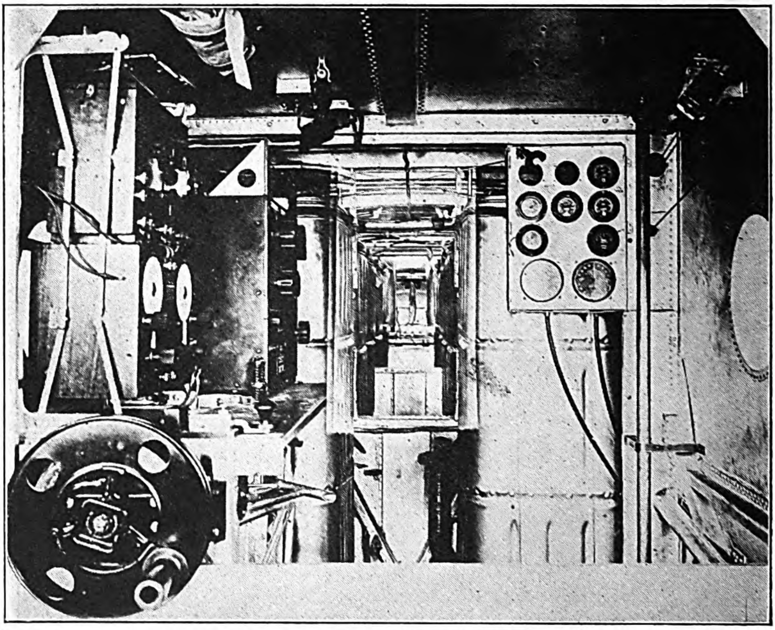

Fig. 41—Radio Installation in Commander Rogers' PN-9 Flying Boat.

A more refined service installation is shown in Fig. 39, which shows the after portion in a PN-7 flying boat; this photograph was taken in 1924 and shows an improved type of compass coil with a series-parallel switch in its center. Note the transposed wiring along the top between compass coil and receiver. Fig. 40 shows a service radio installation in a Torpedo plane where space is somewhat more confined than in flying boats. In Fig. 41 is seen the neat radio installation in the PN-9 metal hull flying boat in which Commander Rogers attempted an Hawaiian flight in 1925. Notice especially the large fuel tanks on either side of the narrow passage-way leading forward, and the all-metal construction, which prevented use of a radio compass coil on account of shielding.

An elaborate installation of Naval intermediate and high-frequency equipment is shown in Fig. 42, installed in the Sikorsky S-37. Of especial interest is the retractible swinging generator mount at the left, which enables withdrawal of the generator from the airstream for inspection purposes and to reduce wind resistance when radio is not employed. Uppermost on

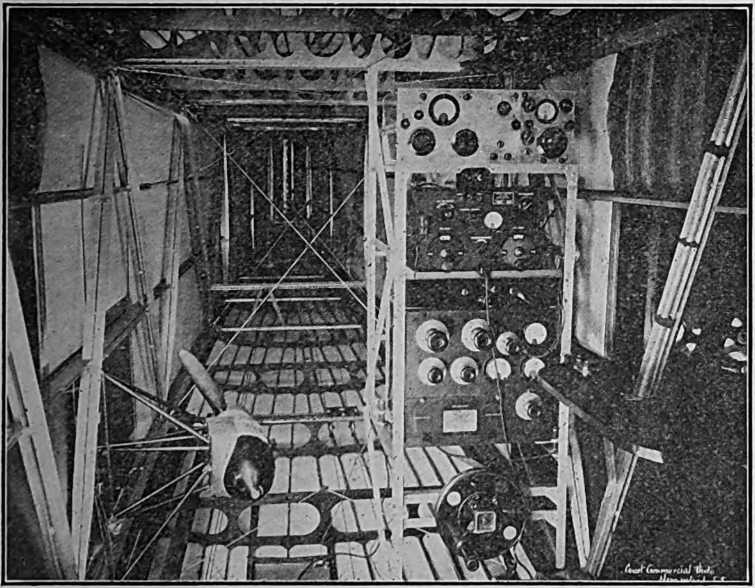

Fig. 42—Radio Installation in Sikorsky S-37 for Fonck's Transatlantic Flight, 1927.

the apparatus rack is shown a high-frequency 10-watt transmitter-receiver built for emergency communication. Below this is a standard intermediate-frequency aircraft receiver, and at the bottom is mounted a standard aircraft transmitter; this transmitter develops about 150 watts into the trailing wire antenna, at intermediate frequencies, employing 400-cycle, self-rectified ACW. The transatlantic flight proposed in this plane last year was indefinitely postponed, so that no opportunity was had thoroughly to test this attractive installation.

K. Directional Radio Devices for Aircraft

The possible applications of directional radio devices to aerial navigation may be broadly classified into three different groups, namely, directional radio receivers on aircraft, directional receiving stations on ground, and directional transmitters on ground.

Radio direction finders installed on aircraft, have in many instances in the past been employed with excellent success and have obvious advantages on large planes operating and navigating independently of extensive ground assistance, as well as for military purposes. The value of such equipment has been shown on various flights, among them, the transatlantic flight of the NC-4 in 1919, the flight of the seaplane *Ne Plus Ultra* from Spain to Argentine in 1925, and on other occasions. Planes equipped with radio direction finders are enabled to obtain bearings for navigational and other purposes from any radio signals intercepted, and thus need not cause additional interference themselves. Rotating coil direction-finding loops have been used on our rigid airships as well as in the earlier Naval flying boats largely of wooden construction. Growing use of metal in airplane construction has rendered such rotating coils inside the hull ineffective in modern planes. By increasing the sensitivity of receivers, however, fairly good results have been obtained with smaller rotating loops projected out of the fuselage, but such loops are objectionable because of their wind resistance. Another method of direction finding in flight employs one or more fixed coils placed around portions of the wing or airplane fuselage and requires

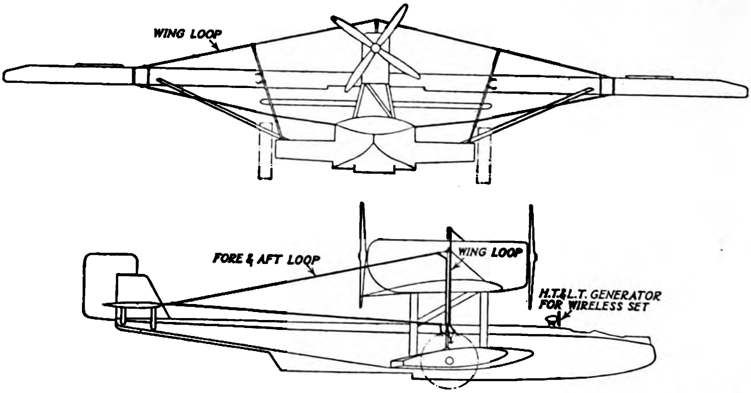

Fig. 43—Arrangement of Fixed Compass Loops on Dornier Flying Boat.

simple loop systems at right angles to each other are placed about the wings or suspended from other convenient parts of the ship, and connect inside the hull to a goniometer in which the field is reproduced and its direction determined with a small rotating search coil which connect to the receiving equipment in the same manner as does a large rotating loop. This latter system has the advantage that bearings on any station may be quickly taken and repeated any number of times without requiring a change in the plane's heading. The Bellini-Tosi system has been used with good success on a number of foreign installations, among them the airship *Norge* and the *Ne Plus Ultra.* Fig. 43 shows the arrangement of the fixed external loops on this flying boat.

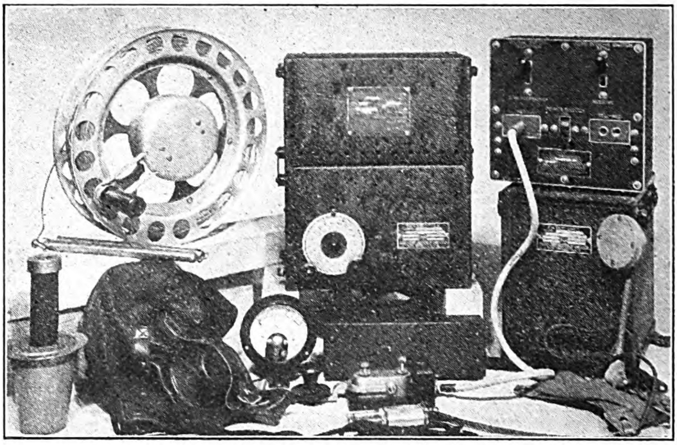

Fig. 45—10-Watt R.C.A. Aircraft Radio Equipment, Model ET-3652.

The placing of direction-finding equipment on aircraft not only requires highly trained personnel among thecrew, but the obtainable accuracy and range of bearings is handicapped by the relatively unfavorable conditions aloft. These drawbacks are avoided if the direction-finding equipment can be placed at favorable locations on the ground, and manned by capable specialists. This system is in general use on the British airways, where planes are enabled at any time to obtain simultaneous bearings from direction-finding stations located at Croydon, Lympne, and Pulham. It is said that the total time required to furnish an airplane with its bearing does not exceed two minutes from the time the plane starts sending its 30-second test signal until it has received its bearing or position.

The directional radio beacon or radio "range," as developed in this country jointly by the Army Air Service and the Bureau of Standards, has the main advantage that nothing but regular receiving equipment is required on the plane, and from the pilot's standpoint is the most easily applied of all methods. The radio beacon is of value not only for navigation on our airways but also has proved its worth as a dependable guide between California and Hawaii; Lieuts. Maitland and Hegenberger received great assistance from these signals, and Arthur Goebel ascribed his Dole flight victory primarily to his constant guidance by the radio beacon.

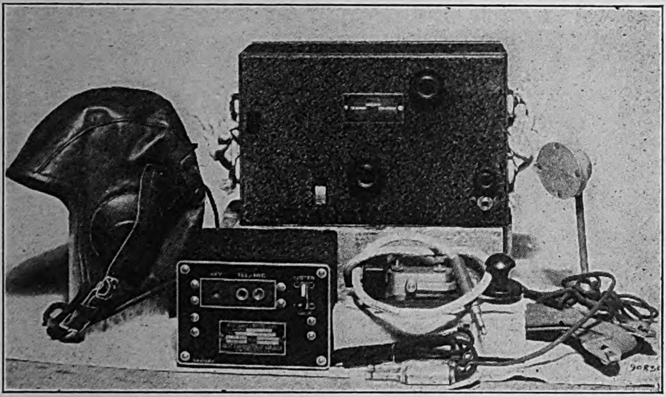

Fig. 46—300-Watt R.C.A. Equipment, Model ET-3654, Showing Receiver and Auxiliaries.

L. Commercial Aircraft Radio Equipment

In Europe there have existed for many years commercial lines of aircraft radio apparatus, the most prominent of which has been developed by the British Marconi Company; their most widely used set, type AD-6, is shown in Fig. 44, installed in a large British passenger plane on the England-Egypt-India air route. The transmitter is rated at 150 watts tube input, and has a transmitting range between 100 and 200 miles, both telegraphy and telephony being provided. The upper portion of the cabinet is occupied by a 5-tube receiver, employing two stages of radio-frequency amplification. An interesting feature of this equipment is provision for full remote control by means of mechanical Bowden cable attachments, so that the equipment may be placed out of the way and operated from the pilot's cockpit.

The Radio Corporation of America has recently placed on the market commercial aircraft radio equipment of three different sizes, giving output ratings of 10, 100, and 300 watts. Both telephone and telegraph communication is provided, in the frequency band between 2250 and 2750 kc. Power supply both for transmitters and receivers is furnished by wind-driven generators, and trailing wire antennas of about 100 feet in length are employed. Fig. 45 shows the components of the 10-watt equipment, except the receiver; the combined weight of the 10-watt equipment is 86½ lbs.

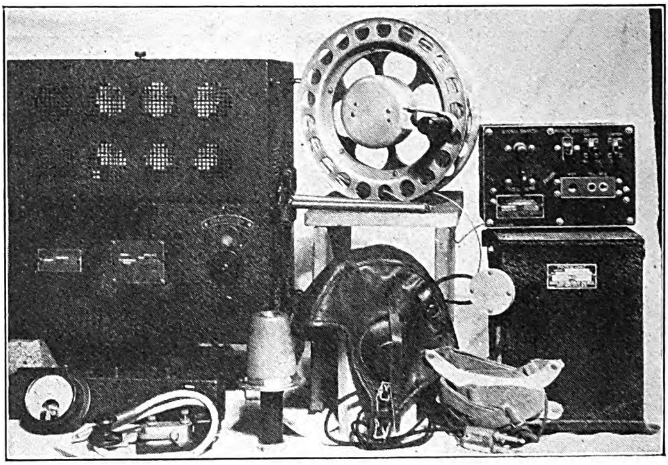

Fig. 47—300-Watt R.C.A. Equipment, Model ET-3654, Transmitter and Related Components.

The 100-watt set, model ET-3653, is designed to cover up to 300 miles by CW telegragh and up to 75 miles or more by telephony; the total weight of this equipment is 133 lbs. A 5-tube receiver with interchangeable coils is a part of this equipment.

The components of the 300-watt equipment model ET-3654, are shown in Figs. 46 and 47. The combined weight of this equipment is 202 lbs., and the range is given as 500 miles for CW and 200 for voice communication. Eight 50-watt tubes model UV-211 are used in the transmitter. The receiver contains five tubes, and is similar to the one employed with the 10-watt equipment. Both the 100- and the 300- watt installations provide for interphone communication between the radio operator and the pilot; in addition, the 300-watt equipment may be operated at will by the pilot by means of an auxiliary remote control unit, shown with the receiver in Fig. 47.

Separate equipment for merely receiving purposes has been developed for commercial airplanes wishing to avail themselves of the weather broadcast and directional beacon service being established on our airways. Frequencies around 300 kc. are employed for these uses.

M. RADIO AIDS FOR OUR AIRWAYS

In accordance with the provision of our Air Commerce Act our national airways and airmail routes are rapidly being provided with adequate radio aids. Among these ground facilities which are being installed under the supervision of the Airways Division, Bureau of Lighthouses, are Radio Stations for intercommunication between airports, ground radio stations for communications with aircraft, directional radio beacons, and low-power marker beacons. There is no doubt that with the provision of adequate ground facilities a great impetus will be given to the equipment of a large portion of our commercial airfleet with modern and efficient types of aircraft radio apparatus, with inevitable gains in the dependability and safety of air transportation.

THE REDUCTION OF ATMOSPHERIC DISTURBANCES*

By

JOHN R. CARSON
(American Telephone and Telegraph Company, New York City)

Summary—In the decade or so during which the problem of eliminating or at least reducing atmospheric disturbances has been given serious and systematic study we have learned, more or less definitely, what we can and cannot do in this direction. For example, we know that there are definite and cannot do in this direction. For example, we know that there are definite limits to what can be accomplished by frequency selection. We know that directional selectivity is of substantial value, particularly when the predominant interference comes from a direction other than that of the desired signal, and we can calculate pretty well the gain to be expected from a given design.

The object of this note is to analyze another arrangement which provides for high-frequency selection plus low-frequency balancing after detection. The broad idea of balancing out the interference is old, but I know of no general analysis of the arrangement. Furthermore the principle of balance has recently acquired fresh interest due to the system disclosed by Armstrong[1] *in which high-frequency selectivity and low-frequency balancing are essential features. Armstrong's scheme is treated in more detail in the latter part of this paper.*

The conclusions of this study are entirely negative, that is, no appreciable gain is to be expected from balancing arrangements. This is quite in agreement with the conclusion drawn over ten years ago by John Mills as a result of a rather extended experimental study made for the Bell System. In fact, as more and more schemes are analyzed and tested, and as the essential nature of the problem is more clearly perceived, we are unavoidably forced to the conclusion that static, like the poor, will always be with us.

I.

IN any theoretical analysis of the static problem we have to face, at the outset, the difficulty inherent in our ignorance of the origin, wave form, and frequency distribution of static. If the problem can be treated as a statistical one this difficulty, as regards practical deductions, can be successfully avoided.[2] When we wish to analyze schemes involving low-frequency balancing after detection, there are serious difficulties in the way of this mode of treatment. Furthermore, it is desirable to have an independent mode of analysis. This is furnished by the following line of reasoning.

Any disturbance, whether signal or atmospheric, can, over any epoch or finite range of time t, be represented by the very general expression[3]

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* Original Manuscript Received by the Institute, April 12, 1928.

[1] PROC. I.R.E., 16, 1, p. 15; Jan., 1928

[2] "Selective Circuits as Static Interference," *Bell System Tech. Jour.*, July, 1925.

$$f(t) \cdot \sin (\omega t + \alpha(t)) \tag{1}$$

with the further restriction that $f(t) \geqq 0$ everywhere and ω is a constant at our disposal. Now let us suppose that (1) represents the disturbance after passing through an efficient selective circuit which confines the transmitted frequencies to those essential to the signal. Then $\omega/2\pi$ can be taken as any frequency inside the transmission band (preferably the carrier frequency of the signal) and f and α will be relatively slowly varying functions. Let us suppose, therefore, that we have a radio receiving system which employs efficient frequency selection before detection. The wave presented to the detector may then be represented by the general expression

$$W = S(t) \cdot \sin (\omega t + \theta(t)) + J(t) \cdot \sin (\omega t + \phi(t)) \tag{2}$$

where the first component represents the desired signal and the second interference. $\omega/2\pi$ is a constant, taken as the signal carrier frequency, and due to the action of frequency selection or filters,

$$\frac{1}{\omega} \frac{d}{dt} \theta\ (t) \text{ and } \frac{1}{\omega} \frac{d}{dt} \phi\ (t)$$

will be small compared with unity. In fact with small error we can write

$$\frac{d}{dt} W = \omega \left\{ S(t) \cdot \cos (\omega t + \theta(t)) + J(t) \cdot \cos (\omega t + \phi(t)) \right\}.$$

No other restrictions are imposed on the amplitudes or phase angles.

In the following we shall limit explicit consideration to radiotelegraph systems, whereby we are permitted to simplify the analysis somewhat by setting $\theta(t) = 0$ in (2). The extension of the analysis to radio telephony, however, presents no essential difficulties and the conclusions of the present study apply without modification to this case also.

II.

In analyzing schemes directed to solving the static problem, false conclusions, unduly favorable to the schemes under consideration, have been drawn, time after time, by reason of the simple failure to compare the specific arrangement under analysis with a standard of reference. In the present discussion our standard of reference will be defined as follows. After the received wave is passed through a band filter, or a selective circuit such that the frequencies present in W, as given by (2), are confined to those essential for signaling purposes, the wave is demodulated or detected by a "homodyne" of carrier frequency $\omega/2\pi$. The detected output is then represented by

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[3] See Appendix to this paper.

$$W' = \tfrac{1}{2}S(t) + \tfrac{1}{2}J(t) \cdot \cos(\phi). \tag{3}$$

Possibly this "reference system" requires a word more of explanation. At the transmitting station let the low-frequency signal $S(t)$ modulate a carrier wave represented by $\sin \omega t$, where $\omega/2\pi$ is the carrier radio frequency. The modulated output is then given by the expression

$$S(t) \cdot \sin \omega t.$$

Now let us suppose that $S(t)$ is representable by a series of sinusoidal terms (in the limit a Fourier's integral); that is, suppose

$$S(t) = \sum a_m \sin(\omega_m t + \theta_m).$$

Then by a well-known trigonometric formula, the modulated output is analyzable into two side bands

$$\tfrac{1}{2}\sum a_m \cos[(\omega - \omega_m)t - \theta_m] - \tfrac{1}{2}\sum a_m \cos[(\omega + \omega_m)t + \theta_m].$$

In one side band the frequency is less than that of the carrier wave; in the other greater.

We now suppose that by means of filters or otherwise one side band is suppressed and only one transmitted; say the lower side band. At the receiving station the unmodulated carrier wave $\sin \omega t$ is restored and it, together with the transmitted side band, impressed on the input circuit of a square-law demodulator. The demodulated output is then

$$\begin{aligned}&\tfrac{1}{2}\sin \omega t \sum a_m \cos[(\omega - \omega_m)t - \theta_m] \\ &= \tfrac{1}{4}\sum a_m \sin[(2\omega - \omega_m)t - \theta_m] + \tfrac{1}{4}\sum a_m \sin(\omega_m t + \theta_m).\end{aligned}$$

The first summation represents double radio-frequency waves, which can easily be suppressed in the low-frequency circuit by selective means. The second summation is simply the original low-frequency signal. This system of transmission, it may be remarked, is employed in transatlantic radio-telephony.

Returning to (3) it will be observed that in our standard of reference, the output is linear in signal and interference amplitudes, S and J, and that the ratio of interference to signal is

$$\frac{J \cos \phi}{S}. \tag{5}$$

Since the phase angle ϕ is entirely arbitrary, the mean value of the ratio is

$$\frac{2}{\pi} \frac{J}{S}. \tag{6}$$

Any scheme, proposed for the solution of the static problem, must, in order to prove in, show a smaller ratio of interference to signal.

In the case of radio-telephony, demodulation by the homodyne principle or its equivalent is essential for high quality. In telegraphy, however, the same requirement is not present. Suppose, therefore, we examine the case of a *square-law* detector. With such a device the low-frequency output is given by

$$\tfrac{1}{2}S^2+\tfrac{1}{2}J^2+SJ \cos (\phi). \tag{7}$$

Since the phase angle ϕ is uncontrollable and, in general, variable, the last term of (7) is equally likely to be positive or negative. For the case of very strong interference, (7) becomes

$$\tfrac{1}{2}S^2+\tfrac{1}{2}J^2 \tag{8}$$

while for weak interference it is

$$\tfrac{1}{2}S^2+SJ \cos (\phi). \tag{9}$$

Comparing with (6), the corresponding expression for our standard of reference, it is seen that, in both cases, the interference ratio is increased by square-law detection.

In the case of radio-telegraphy there is another mode of demodulation, namely *straight-line* or *linear rectification.* There seems to be a more or less prevalent belief that this method possesses advantages over square-law detection, particularly as regards intermodulation between signal and interference. This belief is not justified; in fact it is an erroneous inference from the relative difficulty of analyzing the rectified output of a complex wave, and the very complex character of that output itself. In order to analyze exactly the rectified output, the wave form of the input wave must be exactly specified, this requiring information which is never available in practice. General and qualitative information, sufficient for our purposes, can, however, be deduced as follows.

Returning to the wave, W, as given by (2), as the wave impressed on the ideal straight-line rectifier, this can be written as

$$W = F \cdot \sin(\omega t + \psi(t)) \tag{10}$$

where

$$F = \sqrt{\{S^2 + J^2 + 2SJ \cdot \cos(\phi)\}} \tag{11}$$

and

$$F \cdot \sin \psi = J \sin \phi .$$

The output wave is then given by

$$W' = M \cdot W = M\{S \sin(\omega t) + J \sin(\omega t + \phi)\}, \tag{12}$$

where M is a *modulating wave* defined as *unity* when $\sin(\omega t + \psi) > 0$ and *zero* when $\sin(\omega t + \psi) < 0$.

In the idealized case when the interference is absent, the modulating function of equation (12) is given by

$$M = \tfrac{1}{2} + \sin(\omega t) + \tfrac{1}{3} \sin(3\omega t) + \tfrac{1}{5} \sin(5\omega t) + \cdots \tag{13}$$

Comparing (13) with our standard of reference system it is seen that M differs from the demodulating homodyne $\sin(\omega t)$ only by the presence of the zero frequency and harmonics of the carrier frequency. Theoretically perfect demodulation, however, results.

When, however, the signal wave is complex and in addition interference is also present, the modulating function M and the demodulated output of the rectifier is vastly more complicated. Making, however, certain ideal assumptions—the most favorable possible to the method of demodulation under immediate consideration—it may be shown that the predominant term of the demodulated output is represented approximately by

$$\frac{m}{2}\{S(t)\cos(\psi) + J(t)\cos(\phi - \psi)\}. \tag{14}$$

Here m is itself a variable depending on the relative amplitudes and phases of the signal and interference. In addition the phase angle ψ is an uncontrollable variable. In practice, in addition to the output, as given by (14) there are terms of distorted frequencies which cannot be filtered out in the low-frequency circuit.

However, the analysis of the linear rectifier is much more simply effected by employing the following approximate expression for the demodulated low-frequency output:

$$W' = \tfrac{1}{2}\sqrt{S^2 + J^2 + 2SJ\cos(\phi)} \qquad (15)$$

This is the *envelope* of the high-frequency disturbance and, it will be observed, is proportional to the square root of the output of the square law detector. This formula, while in general approximate, is quite sufficiently accurate for our purposes, and is indeed favorable to the linear rectifier.

III.

The foregoing completes our discussion of the elementary theory of demodulation with particular reference to the simultaneous presence of signal and interference, which is the critical case to be considered. It remains to apply this theory to low-frequency balancing arrangements.

Common to all balancing schemes, the receiving system must include, in addition to the radio-frequency signaling channel, an auxiliary channel of substantially the same frequency-band width and preferably located very close to the former in the frequency scale. This auxiliary channel must be quite sharply selective *against* the signal itself. On the other hand it is supposed to absorb, from atmospherics or other random disturbances, substantially the same amount of energy as the signal channel; indeed this requirement is essential to the theory of the operation of the arrangement.

The received waves, after passing through frequency selective circuits in the two channels are demodulated by separate devices, and the demodulated outputs are differentially combined in a common low-frequency receiver, or receiving circuit.

The idea of the operation of this device is very simple. If the desired signal is alone present the auxiliary channel does not affect reception since, by means of frequency discrimination, it is unresponsive to the signal frequencies. When signal and atmospheric interference, or static, are simultaneously present, the latter is supposed to be balanced out in the low-frequency circuit, since the two high-frequency channels will absorb substantially the same amounts of energy from the interference.

Let us, however, examine the operation of the system in more detail, in the light of the elementary analysis developed above.

Just as in (2), the high-frequency wave in the signal channel, after frequency selection, may be written as

$$W = S(t) \cdot \sin(\omega t) + J(t) \cdot \sin(\omega t + \phi(t)) \tag{15}$$

In the auxiliary channel the corresponding wave is

$$W_1 = J_1(t) \cdot \sin(\omega_1 t + \phi_1(t)) \tag{16}$$

ω_1 may be taken as a constant nearly equal to ω, their difference depending on the frequency separation between channels. The relation between J_1 and J and ϕ_1 and ϕ will depend on the wave form of the interference.

Now let us suppose that the waves, as given by (15) and (16) are demodulated by homodyne generators of frequency $\omega/2\pi$ and $\omega_1/2\pi$. Corresponding then to (3) the demodulated output from the signal channel is

$$\tfrac{1}{2}S(t) + \tfrac{1}{2}J(t)\cos(\phi) \tag{17}$$

while that from the auxiliary channel is

$$\tfrac{1}{2}J_1(t)\cos(\phi_1). \tag{18}$$

Subtracting (18) from (17) the resultant low-frequency output is

$$\tfrac{1}{2}S(t) + \tfrac{1}{2}J(t)\cos(\phi) - \tfrac{1}{2}J_1(t)\cos(\phi_1). \tag{19}$$

Now since the phase angles ϕ and ϕ_1 are both variable and uncontrollable, it follows that the two interference components are equally likely to add or subtract so that no gain by balancing results on the average.

Suppose, however, demodulation in both channels is effected by a square law detector. The differential low-frequency output is then

$$\tfrac{1}{2}S^2 + \tfrac{1}{2}(J^2 - J_1^2) + SJ \cdot \cos(\phi). \tag{20}$$

Let us assume further that $J^2 - J_1^2$ can be made negligibly small, a condition at least partially realizable. The output is then

$$\tfrac{1}{2}S^2 + SJ \cdot \cos(\phi). \tag{21}$$

It follows from (21) that the interference is effectively suppressed *in the absence of the signal.*[4] Comparison with (3), however, shows that, when the signal is present, the interference to signal ratio is just twice that obtainable with our reference standard circuit. Whether or not the suppression of interference in the absence of the signal compensates for the increased interference-to-signal ratio in the presence of the signal is an open question which can only be decided by practical experience. Furthermore the balance obtainable in practice will certainly be far from perfect. It is further to be observed that since homodyne demodulation, or its equivalent, is essential for telephonic signals, no gain by balancing is possible in radio-telephone transmission. An added disadvantage, it may be noted, that attaches to balancing schemes is that the receiving system must be responsive to a frequency range double that required for the usual method of reception.

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[4] See, however, the concluding paragraph of this paper.

If the analysis is extended to the case of straight-line rectifier demodulation by means of (15) the general conclusions are of the same character. In fact we have for the resultant demodulated output

$$\tfrac{1}{2}\sqrt{S^2+J^2+2SJ\cos(\phi)}-\tfrac{1}{2}J_1 \qquad (22)$$

which, in the absence of the signal (spacing interval) becomes

$$\tfrac{1}{2}(J-J_1) \qquad (23)$$

As in the case of the square law detector, the gain results only during the spacing interval and is obtained at the expense of an increased interference ratio during the marking interval.

IV.

The foregoing reasoning will now be applied to the receiving system proposed by Armstrong. His arrangement specifies demodulation by rectification. Its only essential difference from the corresponding balancing scheme discussed above is that in the normal absence of the signal (spacing interval) a wave of slightly different frequency is transmitted to which the auxiliary channel is responsive. Applying (15), we have in the presence of the signal (marking interval) the following expression for the demodulated output

$$\tfrac{1}{2}\sqrt{S^2+J^2 2SJ\cos(\phi)}-\tfrac{1}{2}J_1 \qquad (24)$$

while during the spacing interval, we have

$$\tfrac{1}{2}J-\tfrac{1}{2}\sqrt{S_1{}^2+J_1{}^2+2S_1J_1\cos(\phi_1)}. \qquad (25)$$

Comparing (22) with (24) it is clear that no gain results, in the marking interval, over that procured with an ordinary balance: In fact, during this interval the outputs are identical. In the spacing interval, however, a comparison is quite unfavorable to the Armstrong scheme. For, in the usual balance system, it is theoretically possible to balance out substantially interference in the absence of the signal. In the Armstrong system, however, interference occurring during a spacing interval may result in a false signal, depending on the intensity of the interference, and on uncontrollable, variable phase angles.

There is another feature of the Armstrong arrangement which must be taken into account in comparing it with standard systems. This is that, as compared with non-balancing arrangements, the Armstrong system requires a doubling of the power radiated and a doubling of the receiving frequency band. *By discarding the balancing feature and the spacing wave, it should be possible to transmit by usual circuits, the signal message and a repetition thereof with the same power and the same frequency requirements of the receiving system.* It would be extremely interesting to have a comparison of such a system with that proposed by Armstrong.

In the foregoing discussion the possibility of balance in the absence of signal (spacing interval) was treated optimistically in order that the conclusions of the analysis should be conservative, and for the sake of simplicity. That the balance ordinarily obtainable in the spacing interval, even with the ideal rectifier, is likely to be quite imperfect may be seen by the following discussion of a case of probably frequent occurrence touched on by Englund[5] in his discussion; namely, where the interference consists of two or more overlapping disturbances. To consider this case, the interference in the receiving channel may be written as

$$J \sin(\omega t+\phi)+J' \sin(\omega t+\phi')$$

while in the auxiliary balancing channel it is

$$J_1 \sin(\omega t+\phi_1)+J_1' \sin(\omega t+\phi_1').$$

The differential rectified output is then approximately

$$\sqrt{J_2+J'^2+2JJ' \cos(\phi-\phi')}-\sqrt{J_1^2+J_1'^2+2J_1J_1' \cos(\phi_1-\phi_1')}.$$

Now even if we grant the possibility of balancing completely the terms $(J-J_1)$ and $(J'-J_1')$ the presence of the uncontrollable variable phase term in the preceding makes it clear that the balance will, in general, be imperfect for the case of overlapping disturbances. As stated above, it is reasonable to suppose that this case is of frequent occurrence.

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[5] Proc. I. R. E., **16**, 1, p. 27; Jan. 1928.

APPENDIX

Formula (1) can be established as follows: Any disturbance, supposed to exist only over a finite interval of time, t, can be formulated as the Fourier integral

$$I=\int_0^\infty F(\lambda)\sin[\lambda t+\theta(\lambda)]d\lambda.$$

Now write $\lambda=\omega+\mu$ where ω is a constant at our disposal; we get

$$\begin{aligned} I=\sin\omega t&\int_{-\omega}^{\infty}F(\mu+\omega)\cdot\cos\,[\mu t+\theta(\mu+\omega)]d\mu\\ &+\cos\omega t\int_{-\omega}^{\infty}F(\mu+\omega)\cdot\sin[\mu t+\theta(\mu+\omega)]d\mu\\ &=I_c\sin\omega t+I_s\cos\omega t\\ &=f(t)\sin\,[\omega t+\alpha(t)]\end{aligned}$$

where

$$f^2(t)=I_c{}^2+I_s{}^2$$
$$\cos\alpha(t)=I_c/f(t).$$

The preceding analysis is entirely formal; its practical significance enters in only when we suppose that the disturbance has passed through an efficient selective network which confines, more or less efficiently, the transmitted frequencies to a finite band, say $\omega_1\leqq\omega\leqq\omega_2$. In this case

$$I=\int_{\omega_1}^{\omega_2}F(\lambda)\sin\,[\lambda t+\theta(\lambda)]d\lambda.$$

If we now write $\lambda=\omega+\mu$ when ω lies within the transmitted band of frequencies, the same analysis shows that if ω is large compared with $\omega_2-\omega_1$, $f(t)$ and $\alpha(t)$ are relatively slowly varying functions.

As stated, the preceding analysis applies rigorously to all types of disturbances. In the case of a radiotelegraph signal, however, it is convenient and permissible to start with the approximate expression

$$S(t)\cdot\sin\omega t$$

where $\omega/2\pi$ is the carrier frequency and $S(t)$ a low-frequency function $\geqq 0$.

THERMOSTAT DESIGN FOR FREQUENCY STANDARDS*

By

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THE means usually employed for maintaining constant the temperature of a body employs a device, responsive to small variations from a given temperature, which automatically varies the rate of heating whenever the temperature departs from this prescribed value. For the operation of such controlling means it is necessary for the temperature of the responding device to vary between a value slightly above its normal operating value and one slightly below, since it is only through such variation that any control can be effected. Of necessity the temperature of the material immediately surrounding the responding element fluctuates even more, whether it be a fluid bath or any other medium.

The body which is to be maintained at constant temperature and the controlling element often bear the same thermal relation to the heater. If this is the case, the same temperature variations that operate the responding element are applied to the region it is desired to control. Sometimes the responding element is mounted within the controlled region, which from this standpoint is an undesirable arrangement. In order to reduce the variation in such cases, it is necessary to employ especially sensitive responding elements and to make their heat capacity small in comparison with that of the controlled object.

The same end may be attained by so designing the thermal system that the variations reaching the object to be controlled are materially reduced below those necessarily existing at the responding element. To accomplish this a layer of material, especially chosen for the property of attenuating temperature variations, is interposed between the object to be controlled and the region about the responding element.

The resulting arrangement of the thermostat wall is shown in Fig. 1. The four essential layers consist of a layer of thermal insulation, a heater as well distributed as possible immediately

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within the insulating wall, a distributing layer for equalizing temperatures parallel to the surface, and what we shall call an "attenuating layer" for reducing the effects of the time variations of temperature which exist in the distributing layer.

The insulating and heating layers need no comment except that it is well to keep the simplest possible symmetry in order to aid in the distribution of heat. From this standpoint there is a real advantage in using a cylindrical shape as indicated by the sketch Fig. 1.

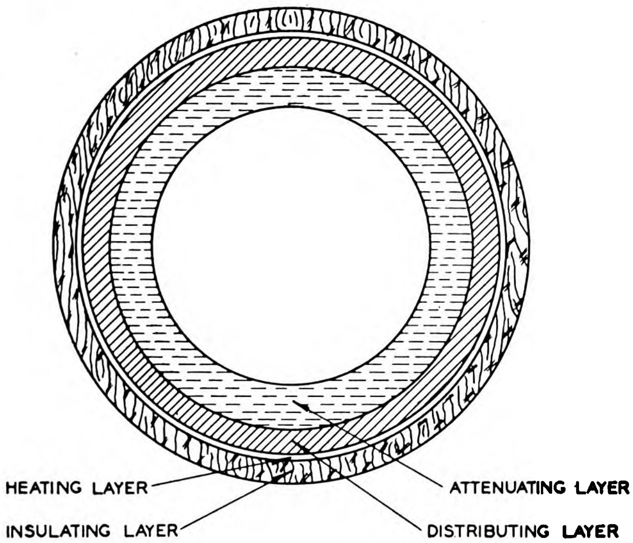

Fig. 1.

The distributing layer should have high thermal conductivity parallel to the surface and a low heat capacity, in order to equalize the temperature over the surface completely and quickly and to permit the responding element to operate as frequently as possible. A stirred bath of light oil has both of these properties and is satisfactory where space and conditions permit the use of a liquid, but a wall of highly conducting metal with low density, like aluminum, is more desirable from most standpoints and is especially effective if laminated, with the layers separated by thin insulation, like paper.

The attenuating layer should be made of material having large heat capacity and low conductivity, the effectiveness depending on the ratio DH/C, where D is the density, H is the specific heat, and C is the conductivity. If these constants for any substance are known, the temperature variations at any point in an infinite wall (or any wall of large radius) can be expressed as

$$T = T_0 e^{-\sqrt{\frac{DH\omega}{2C}}\,l}$$

where T_0 is the applied temperature variation and T is the variation at any point, distance l from the surface. The constant ω is 2π times the frequency of thermostat operation. In Table I a list of ordinary substances is given with the attenuation constant $\sqrt{\frac{DH}{2C}}$ computed from constants found in various tables.

TABLE I

ATTENUATION CONSTANT $\sqrt{\frac{DH}{2C}}$ FOR A FEW ORDINARY SUBSTANCES

Substance	Density-D	Specific Heat-H	Conductivity-C	$\sqrt{\frac{DH}{2C}}$
Asbestos	3.0	0.20	0.00019	40.
Wood (Cedar ⊥ grain)	0.5	0.42	0.00009	34.
Vulcanite	1.8	0.33	0.00042	27.
Glycerine	1.26	0.54	0.0006	25.
Turpentine	0.87	0.42	0.0003	25.
Water	1.00	1.00	0.001	22.
Paraffin	0.8	0.69	0.31	22.
Rubber	0.9	0.4	0.00045	20.
Sulphur	2.0	0.18	0.00045	20.
Alcohol	0.79	0.5	0.0005	20.
Phenol Fibre	1.35	0.35	0.0006	20.
Wool Felt	0.32	0.3	0.00013	19.
Cellotex	0.26	0.26	0.00014	16.
Silocel (Powder)	0.22	0.24	0.00013	14.
Porcelain	2.3	0.26	0.0025	10.8
Glass	3.0	0.15	0.002	10.6
Silica	2.2	0.19	0.002	10.2
Mercury	13.6	0.03	0.02	3.2
Air	0.0013	0.24	0.00005	1.8
Iron	7.86	0.1	0.14	1.7
Lead	11.3	0.03	0.092	1.4
Aluminum	2.7	0.22	0.5	0.77
Hydrogen	0.000090	3.4	0.0003	0.71
Copper	8.9	0.09	1.0	0.63
Silver	10.5	0.055	1.0	0.54

It has been found that the best attenuating property is obtained when the constants in the attenuating wall are tapered so that on the inside the heat capacity is a maximum and on the outside the thermal conductivity is a minimum. This can be accomplished, in a laminated structure using two materials, by using laminae of varying thickness so arranged that one material predominates on one side of the wall and the other material predominates on the other side. The effectiveness of the attenuating wall increases rapidly with frequency of thermostat operation

indicated between the layers. Such discontinuities occur at each boundary and assist materially in reducing variations in temperature, especially if a laminated structure is used.

As an example of what can be gained by the use of an attenuating layer, consider an oil bath in which are immersed two containers, as shown at *A* and *B* in Fig. 3, one being in direct contact with the bath and the other separated from it by an attenuating wall, as described, one centimeter thick. The temperature variations applied to container *A* are, of course, those of the bath itself. Those applied to container *B* are considerably reduced by the attenuating layer as indicated by the table in Fig. 3. When the period of thermostat operation is one minute or less, and when an attenuating layer with tapered constants as previously described is used, the temperature variation is reduced in the ratio of at least 10,000 to one.

Discussion on

RECENT DEVELOPMENT IN LOW POWER AND BROADCAST TRANSMITTERS* (I. F. BYRNES)

F. M. Ryan†: Mr. Byrnes is to be complimented on having given some quantitative data on harmonic radiation from radio transmitters. The reduction of such radiation is an important problem and has been given far too little attention in the literature.

It is noted that Mr. Byrnes expresses the magnitude of harmonics in terms of the ratio of the power of the fundamental to that of the harmonic. The limiting value of this factor is, of course, what the design engineer requires in determining the attenuation he must provide to currents of these unwanted frequencies. It should be kept in mind, however, that the interference caused by harmonic radiation is determined solely by the absolute value of the radiated power. For example, with a fundamental-to-harmonic power of 30,000 to 1 in the case of a 2000-watt transmitter the same degree of interference will be experienced as with a fundamental-to-harmonic ratio of 3000 to 1 in a 200-watt transmitter, each producing a harmonic field strength of about 1800 microvolts per meter at a distance of one mile.

It is interesting to note that in order to limit the radiation of the second harmonic to an absolute value equivalent to the figure mentioned by Mr. Byrnes (0.067 watt) it would be necessary in the case of a 50-kw radio transmitter to attain a fundamental-to-harmonic power ratio of 750,000 to 1. With this degree of harmonic reduction, the field strength at one mile would again be 1800 microvolts per meter. As a matter of fact, tests of the Western Electric 50-kw Radio Broadcasting Transmitter made at Whippany, New Jersey, show a considerably higher degree of harmonic reduction, the field strength of the harmonic of the greatest amplitude being less than 500 microvolts per meter at one mile distance from the station with a fundamental field strength of more than 1,500,000 microvolts per meter. This corresponds to a ratio of fundamental power-to-harmonic power of more than 10,000,000 to 1 or 70 TU.

I. F. Byrnes‡: The reduction of harmonic output and the best method expressing the magnitude of harmonics from transmitting sets are being given considerable attention at the present time. While it is true that the interference caused in a particular receiver at a certain distance from the transmitter is determined by the absolute value of harmonic power, unless a ratio is used to express the difference between the fundamental and some harmonic, it is difficult to compare quickly the merit of transmitters of different power outputs. It is also true that with a given type of transmitting circuit, having the same overall characteristics for different output ratings, the higher powered set is certain to produce harmonics of greater amplitude.

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‡ Radio Engineering Dept., General Electric Co., Schenectady, N. Y.

In the case of broadcast transmitters, it should be realized that the more remote location of high power sets automatically mitigates disturbance due to harmonics, with respect to the number of people annoyed by such harmonics. For example, a 5-kw broadcasting transmitter located in or near a large city and operating on the lower broadcasting frequencies might cause considerable interference on its second harmonic. If we moved this station to a point 20 or 30 miles from the city and increased its power to, say 50 kw, the absolute value of the harmonic may be increased many times without causing any more interference than in the first case, to the same population group.

The design of transmitting sets for commercial or military application presents more difficulty with respect to harmonic reduction than in the case of broadcasting or fixed land stations. In many commercial transmitters which cover a wide frequency band, the frequency we may wish to suppress under one condition is the frequency on which maximum radiation is desired under some other condition. This of course makes the design of output filters a difficult problem. It is fortunate that broadcasting transmitter design permits us to use a fixed transmitting frequency and carefully designed low-pass filters, traps, etc., in the various circuits.

BOOK REVIEW

The World of Atoms. . . By Arthur Haas, translated by Horace S. Uhler. Published by D. Van Nostrand Company, New York, 1928. pp. xi+139, 31 illustrations. Price $3.00.

This little book is decidedly to be recommended to all who, having some knowledge of elementary physics, wish to inform themselves concerning the remarkable advances in atomic theory that the past few years have witnessed. In many respects it resembles the same author's earlier book, "The New Physics," but it is considerably more comprehensive, and traces the subject down to the latter part of 1927. It is based upon a course of lectures given in Vienna. Concerning these lectures the preface states that "they were intended to present to a lay public the achievements of modern atomic physics in as brief and yet thorough a manner as possible, and at the same time in an easily understandable form." The titles of the ten chapters will serve to indicate the nature of the subject-matter: Matter and Electricity; The Building-Stones of Atoms; Light-Quanta; Spectra and Energy Levels; The Elements; The Atom as a Planetary System; Molecules; Radioactivity; Transformations of the Elements; Wave Mechanics of the Atom.

In each chapter there is a clear and logical development of the subject from simple phenomena to the more advanced ideas of the present day. The last chapter is entirely new, having been written especially for the American edition. In it a brave attempt is made to present in popular style the essence of the new wave-mechanics. Considering the difficulty of the subject and the brevity of the treatment, as much success has been achieved as could be expected. In various other parts of the book as well the lay reader will probably wish that the ideas had been more fully explained.

In so comprehensive a treatment complete accuracy can hardly be expected. The reviewer is of the opinion that the discovery of the cathode rays, p. 13, should be attributed to Plücker rather than to Hittorf. On p. 86 is the statement that "crystals represent giant molecules." The present view, however, is that while this is in general true of inorganic crystals, still in many organic crystals there is evidence of the separate existence of individual molecules within the crystal.

Even though it is to be expected that in a book of this type the names of many investigators must necessarily be omitted, still surely in referring to the cathode-ray parabolas the author should have mentioned Sir J. J. Thomson as well as Aston.

The book is beautifully printed and well illustrated. A high tribute should be paid to Professor Uhler for the extraordinarily good translation into English. Professor Haas's large circle of friends in America will be still further increased by this latest contribution of his to the literature of popular science.

W. G. CADY.

BIBLIOGRAPHY ON AIRCRAFT RADIO*

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The rapid development of commercial aviation has brought increased interest in the possibilities of radio as an aid to air navigation. In order that persons working in this field may have a means of ready reference to the work already done, the references on file in the Bureau of Standards have been compiled into this bibliography. These references have been in part published in the monthly reference lists.[1] They give a fairly complete bibliography on the subject up to June, 1928. As a revision of this bibliography may be published at a later time the Bureau of Standards would be glad to receive any criticisms, corrections, or additions to this list. The references are classified by subjects and under each subject the references are in chronological order. The classification is according to "A Decimal Classification of Radio Subjects—An Extension of the Dewey System,"[2] Bureau of Standards Circular No. 138. The part of the general classification of radio subjects used here is as follows:

R520	Aircraft
R520.1	Research
R520.3	Terminology and symbols
R520.4	Lectures
R520.5	Publications
R520.6	Societies
R520.7	Education and Training
R520.9	Historical
R521	Receiving on aircraft
R522	Transmitting from aircraft
R523	Receiving from aircraft
R524	Transmitting to aircraft
R525	Antennas

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[1] A monthly list of references to articles on radio subjects appearing in periodicals has been published for several years by the Bureau of Standards in the Radio Service Bulletin, a publication of the Department of Commerce; Subscription 25 cents per year from the Superintendent of Documents, Government Printing Office, Washington, D.C. Publication of these monthly lists in the Proceedings has begun recently.

[2] Copies of this circular may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D.C.

R526 Radio as navigation aid
R526.1 Beacon systems
R526.2 Direction finders
R526.3 Field localizers
R526.4 Altimeters

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R210 Horton, J. W. and Marrison, W. A. Precision determination of frequency. PROC. I.R.E., **16,** pp. 137–54; February, 1928.

(Carefully controlled piezo oscillator used as a fundamental frequency standard of high precision. Method of direct comparison with time standard described.)

R210 Braillard, R. and Divoire, E. How broadcasting wavelengths are checked. *Wireless World & Radio Review,* **22,** pp. 219–22; Feb. 29, 1928.

(Station at Brussels for checking European stations' frequency—Description of apparatus used includes heterodyne frequency meters and multivibrator.)

R214 Jouaust, R. Le quartz piezo-electrique comme etalon de frequence (piezoelectric quartz as standard of frequency). *L'Onde Electrique,* **6,** pp. 513–32; Nov., 1927.

(Development and theory of piezo oscillators; precautions when used as standard of frequency.)

R214 Harrison, J. R. Piezoelectric resonance and oscillatory phenomena with flexural vibrations in quartz plates. PROC. I.R.E., **15,** pp. 1040–54; December, 1927.

(Experimental and mathematical discussion.)

R214 Jouaust, R. Le quartz piezo-electrique comme etalon de frequence (The piezo oscillator as a standard of frequency). *L'Onde Electrique,* **6,** pp. 580–88; December, 1927.

(Experiments of Tawill and uses of quartz as oscillators.)

R214 Jammet, J. Stabilisateurs de frequence piezoelectriques pour emmeteurs d'ondescourtes (Piezoelectric frequency stabilizers for short wave transmitters). *L'Onde Electrique*, **7**, pp. 5–20; January, 1928.

(Description of a 50-meter crystal-controlled transmitting set showing method of mounting the crystal.)

R214 Cobbold, G. W. A., and Underdown, A. E. Some practical applications of quartz resonators. *Experimental Wireless* (London), **5**, pp. 215–219; April, 1928.

(Advantages of quartz crystals as standards of frequency discussed.)

R214 Crossley, A. Modes of vibration in piezoelectric crystals. PROC. I.R.E., **16**, pp. 416–23; April, 1928.

(Nodes and antinodes on surface of oscillating crystals. Symmetrical arrangement of nodal points permits study of modes of vibration in the crystal. Formulas given for determination of velocity of sound through quartz.)

R214 Watanabe, Y. Der piezoelektrische Resonator in Hochfrequenzschwingungskreisen (The piezoelectric resonator in high frequency oscillation circuits). *Elektrische-Nachrichten Technik*, **5**, pp. 45–64; Feb., 1928.

(Modes of vibration of quartz crystals. Theory of crystals.)

R214 Giebe, E. and Scheibe, A. Piezoelektrische Kristalle als Frequenznormale (Piezoelectric crystals as frequency standards). *Elektrische-Nachrichten Technik*, **5**, pp. 65–82; Feb., 1928.

(Demonstration of modes of vibration in a quartz ring resonator.)

R230 Coil calculations. *Wireless World & Radio Rev.*, **22**, pp. 394–95; Apr. 11, 1928.

(Design data for coils having 2000 to 3000 microhenries inductance.)

R250 Hazel, H. C. A new method for the calibration of ammeters at radio frequencies. PROC. I.R.E., **16**, pp. 70–74; January, 1928.

(Construction and operation of an electron tube designed for use in the measurement of radio frequency currents.)

R251 Nyman, A. Condenser shunt for measurement of high frequency currents of large magnitude. PROC. I.R.E., **16**, pp. 208–217; Feb., 1928.

(Construction of ammeter for high frequency currents consisting of large fixed condenser in parallel with small condenser. Current through small condenser measured by thermoammeter.)

R261 Jansky, C. M. Jr. and Feldman, C. B. A two-range vacuum tube voltmeter. *Jnl.* A. I. E. E., **47**, pp. 126–32; Feb., 1928.

(Design and operation of vacuum-tube voltmeter for 2 overlapping ranges of voltages using a single battery—Effect of wave form and elimination of its effect discussed.)

R261 Medlam, W. B. and Oschwald, U. A. Further notes on the reflex voltmeter. *Experimental Wireless* (London), **5**, pp. 56–60; Feb., 1928.

(Description of electron-tube voltmeter with 80 per cent of scale linear and range of 50 volts. Curves show performance.)

R270 Barfield, R. H. The attenuation of wireless waves over land. *Experimental Wireless* (London), **4**, pp. 25–30; January, 1928.

Also in *Wireless World & Radio Review*, **22**, pp. 2–6; January 4, 1928.

(Intensity measurements with portable apparatus on transmissions of 2LO (London) giving map of distribution.)

R270 Hollingworth, J. A new universal long-wave radio intensity measuring set. *Jnl. Scientific Instruments* (London), **5**, pp. 1–9; January, 1928.

(New apparatus designed for observation of polarization of radio waves. Intensities of the two components of polarization and their phase relations relative to the ground wave can be measured in quick succession. Constructional details, theory and method of operation are given.

R270 McIlwain, K. and Thompson, W. S. A radio field strength survey of Philadelphia. Proc. I.R.E., **16**, pp. 181–92; Feb., 1928.

(Field strength measurements of WFI broadcasting station in Philadelphia. A Radio field strength contour map of Philadelphia is given.)

R270 Barfield, R. H. The attenuation of wireless waves over land. *Jnl. I. E. E.* (London), **66**, pp. 204–18; Feb., 1928.

(Intensity measurements on transmission of 2LO (London) giving results of investigation which showed greater attenuation than that expected from the Sommerfeld theory.)

R270 Sreenivasan, K. A short survey of some methods of radio signal measurement. *Experimental Wireless* (London), **5**, pp. 205–10; April, 1928.

(Describes various methods now in use for signal intensity measuring apparatus.)

R 300. Radio Apparatus and Equipment

R320 Chireix, H. and Villem, R. Compensation des courantes induits entre antennes emettrices voisines (Compensation of induced currents between nearby transmitting antennas). *Rev. Gen. de l'electricite*, **23**, pp. 523–36; Mar. 24, 1928.

(Analysis of qualitative phenomenon concerning mutual induction between neighboring antennas.)

R323 Le Marquand, H. Sous-marins et ondes Hertziennes (submarines and Hertz waves). *QST Francais et Radio-electricite Reunis*, **8**, pp. 40–42; December, 1927.

(Description of apparatus used on submarines including experiments with different types of antennas.)

R325.1 Kenn, R. Wireless direction finding and directional reception (book, 2nd edition). Publishers, Iliffe & Sons, Ltd., London, 1927. Price 21 shillings. Abstract in *Wireless World & Radio Rev.*, **22**, p. 10, January 4, 1928.

(Methods and apparatus used in direction finding.)

R330 Forstmann and Schramm. Die Elektronenrohre (book). Publishers, Schmidt & Co., Berlin, 1927. Reviewed in *Experimental Wireless* (London), **4**, p. 760; Dec., 1927.

(Treatise on electron tubes.)

R330 The UX250-CX350 tube. *QST*, **12**, p. 36, Apr. 1928.

(Characteristics for this type tube—25-watt power.)

R330 The Frenotron valve—A Vienna novelty. *Experimental Wireless* (London), **5**, p. 214; April, 1928.

(New type of detector to be used as a stabilizer.)

R330 Hanna, C. R., Sutherlin, L., and Upp, C. B. Development of a new power amplifier tube. Proc. I.R.E., **16**, pp. 462–75; April, 1928.

(Description of the development of the power tube UX-250. Properties and limitations as well as design information applicable to these tubes and their circuits are discussed.)

R330 Wheeler, H. A. Measurement of vacuum-tube capacity by a transformer balance. Proc. I.R.E., **16**, pp. 476–81; Apr., 1928.

(Description of complete portable apparatus for measurement of direct capacities of electron tubes in laboratory or factory testing. Tube capacity compared with standard variable condenser by means of transformer-balance (Neutrodyne) circuit.)

R330 Walsh, L. A direct-capacity bridge for vacuum-tube measurements. Proc. I.R.E., **16**, pp. 482–86; Apr., 1928.

(Bridge is described which permits the measurement at a single setting of a capacity associated with other capacities in a system having more than 2 terminals, such as grid-plate capacity of an electron tube.)

R330 Hoch, E. T. A bridge method for the measurement of interelectrode admittance in vacuum tubes. Proc. I.R.E., **16**, pp. 487–93; Apr., 1928.

(Description of Colpitts-Campbell bridge as applied to measurement of direct admittances in electron tubes. Bridge circuit for measurement of direct capacity and conductance given. Data given on several tubes.)

R331 Simon, H. Einiges uber Empfangerrohren (something on receiving tubes). *Telefunken Zeitung*, **11**, pp. 38–50; October, 1927.

(Construction and description of thoriated filament electron tubes.)

R334 The Cossar screened valve. *Wireless World & Radio Rev.*, **21**, pp. 817–18; Dec. 21, 1927.

(Characteristic curves and practical hints for best conditions of operation.)

R334 Warner, J. C. Some characteristics and applications of 4-electrode tubes. Proc. I.R.E., **16**, pp. 424–46; April, 1928.

(Theory, construction and use of 4-electrode tubes.)

R334 Hall, N. R. Characteristic curves of the 4-electrode valve. *Experimental Wireless* (London), **5**, pp. 198–200; Apr., 1928.

(Curves given for British 4-electrode electron tubes.)

R340 The power factor and capacity of the electrodes and base of triode valves with special reference to their use in thermionic voltmeters. *Experimental Wireless* (London), **5**, p. 16; January, 1928.

(Table giving capacity and power factor of several types of tube bases.)

R341 Kuhlman, J. H. and Barton, J. P. The vacuum-tube rectifier *Jnl.* A. I. E. E., **47**, pp. 17–24; January, 1928.

(Design of rectifier for use as B power supply; determination values of inductance and capacity for filter circuit—uses electron-tube voltmeter.)

R341 Loftin, E. H. and White, S. Y. Direct-coupled detector and amplifiers with automatic grid bias. Proc. I.R.E., **16**, pp. 281–86; March, 1928.

(System for direct coupling of electron tubes to give composite detection and amplification which is free from electrical and acoustical feed-back effects.)

R342 Vreeland, F. K. On the distortionless reception of a modulated wave and its relation to selectivity. Proc. I.R.E., 16, pp. 255–80; Mar., 1928.

(Description of an amplifier giving uniform amplification over entire width of band also description of band selector having rectangular frequency characteristic. Various applications of these to broadcast reception.)

R342 Marcus, A. A triode amplifier for small direct currents (abstract). *Physical Rev.*, 31, p. 302; Feb., 1928.

(Measurement of ionization currents and piezo-electric currents of magnitudes considerably below range of galvanometer by use of electron-tube amplifier.

R342 Terman, F. E. The inverted vacuum tube, a voltage reducing power amplifier. Proc. I.R.E., 16, pp. 447–61; Apr., 1928.

(By the interchange of functions of grid and plate of an electron tube a voltage reducing power amplifier is obtained. Theory given.)

R342.15 Muller and von Ardenne. Transformatoren—Verstarker. (transformers—amplifiers) (book). Publishers, Schmidt & Co., Berlin, 1927. Reviewed in *Experimental Wireless* (London), 4, p. 760; Dec., 1927.

(Treatise on radio transformers and amplifiers.)

R342.15 Osborn, B. K. Data on the voltage amplification of r.f. transformers. *Radio Engineering*, 8, pp. 24–25; Apr., 1928.

(Apparatus used and operations involved for testing of r.f. transformers.)

R342.6 Runge, W. Der abgestimmte Hochfrequenzverstarker (the tuned radio frequency amplifiers). *Telefunken Zeitung*, 11, pp. 50–63; October, 1927.

(Theory of radio-frequency amplifiers with special reference to neutralization.)

R342.6 Inglis, C. C. Good quality in high-frequency amplifiers. *Experimental Wireless* (London), 5, pp. 132–33; Mar., 1928.

(Calculation of effect of high impedance electron tubes on the sharpness of resonance of tuned circuits connected to them.)

R344 Prince, D. C. and Vogdes, F. B. Vacuum tubes as oscillation generators. General Electric *Rev.*, 31, pp. 97–98; Feb., 1928.

(Design of simpler electron-tube circuits. Difference between Hartley and Colpitts circuits.)

R344 Heim, W. Dispersionmessungen im Gebiete kurzer elektrischen wellen (Dispersion measurements in direction of short electric waves). *Zeitschrift für Hochfrequenztechnik*, 30, pp. 176–83; Dec., 1927.

(Description of electron-tube generator and application to short wave measurements.)

R344 Reed, M. Parasitic oscillations in the case of a tuned-anode oscillator. *Experimental Wireless* (London), 5, pp. 135–47; March, 1928.

(Conditions under which parasitic oscillations may occur and discussion of their cause.)

R344 Prince, D. C. and Vogdes, F. B. Vacuum tubes as oscillation generators. General Electric *Rev.*, 31, pp. 147–52; March, 1928.

(Special considerations of design and operation of generating circuits.)

R344.3 Hollman, H. E. Transmitting on a wavelength of 3/4 of a meter. *Radio News*, 9, pp. 1143–45; April, 1928.

(Description of apparatus used by author for transmission of telephony on wavelength of 75 cm.)

R348 Knowles, D. D. The grid-glow tube relay. *Electric Jnl.*, 26, pp. 176–78; April, 1928.

(Description of a relay which has an amplification factor of 10^5 under certain conditions. Gives miscellaneous applications of this relay.)

R351 Rodwin, G. and Smith, T. A. A radio frequency oscillator for receiver investigations. PROC. I.R.E., 16, pp. 155–65; Feb., 1928.

(Description of modulated radio frequency oscillator with means for obtaining radio frequency outputs for measuring purposes.)

R360 Wheeler, H. A. Automatic volume control for radio receiving sets. PROC. I.R.E., 16, pp. 30–39; January, 1928.

(Uses rectified carrier voltage to adjust grid bias of radio-frequency amplifier tubes.)

R374 Schleede, A. and Biggish, H. Zusammenfassender Bericht—Der Kristalldetektor (Joint report—Crystal detectors). *Zeitschrift für Hochfrequenztechnik*, 30, pp. 190–93; December, 1927.

(Theory of natural and manufactured crystals for use as radio detectors.)

R376 Wente, E. C. and Thuras, A. L. A high efficiency receiver for a horn type loudspeaker of large power capacity. Bell System Technical *Jnl.*, 7, pp. 140–53; Jan., 1928.

(Telephone receiver of moving coil type. Used in Movietone and Vitaphone.)

R376.3 Story of the hornless loudspeaker. *Wireless World & Radio Rev.*, 21, pp. 806–10; Dec. 21, 1927.

(Historical survey 1879 to 1927 of development of free-edge, fixed-edge, and inertia-controlled cone loudspeakers.)

R376.3 Meyer, E. Uber die nichtlineare Verzerrung von Lautsprechern und Fernhoren (On the non-linear distortion of loudspeakers and receivers). *Elektrotechnische-Nachrichten Technik*, 4, pp. 509–15; December, 1927.

(Results of tests carried out by means of a sound compensation device which measures the strength of the overtones in proportion to the strength of fundamental tones.)

R376.3 Hanna, C. R. Abridgment of loudspeakers of high efficiency and load capacity. *Jnl.* A. I. E. E. 47, pp. 253–57; April, 1928.

(Design of high quality horntype loudspeaker with moving coil driving elements considered. Calculation of efficiency and maximum output capacity from this type of speaker.)

R381 Smith, B. E. The design of variable condensers for high voltage operation. *QST*, 12, pp. 49–51; March, 1928.

(Qualifications of condensers to withstand high voltages. Use in large transmitting stations.)

R381 Smith, B. E. The relation of condenser ratings to filter design. *Radio Engineering*, 8, pp. 33–34; Apr., 1928.

(Explanation of voltage regulation in power supply units and its effect on the filter condensers.)

R383 Hitchcock, R. C. Designing fixed resistors. *QST*, 12, pp. 29–32; April, 1928.

(Chart is given for rapid calculation of resistances of various kinds and sizes of wire together with safe current rating of a resistor.)

R384.1 Colebrook, F. M. Description of a valve wavemeter with a range of 10 to 20,000 meters. *Experimental Wireless* (London), **4**, pp. 721–25; Dec., 1927.

(Precision frequency meter independent of tube variations.)

R384.1 Reed, M. The suppression of parasitic oscillations in valve circuits. *Experimental Wireless* (London), **4**, pp. 725–32; Dec., 1927.

(Application to heterodyne frequency meter.)

R384.1 Griffiths, W. H. F. Notes on the accuracy of variable air condensers for wavemeters. *Experimental Wireless* (London), **4**, pp. 754–57; Dec. 1927.

(Requirements for a condenser for use in precision frequency meter.)

R384.1 Griffiths, W. H. F. The accuracy and calibration permanence of variable air condensers for precision wavemeters. *Experimental Wireless* (London), **5**, pp. 63–74; February, 1928.

(Discussion of condenser construction as to errors introduced by stray capacities and their elimination.)

R385.5 Jakowleff, A. J. Die Berechnung der akustischer Eigenschaften des Kondensatormikrophons (calculation of the acoustical qualities of a condenser microphone). *Zeitschrift für Hochfrequenztechnik*, **30**, p. 151; Nov., 1927.

(Theory and operation of the condenser microphone.)

R386 Replogle, D. E. and Millen, J. The final capacity in a 2-section low frequency filter. *QST*, **12**, pp. 36–38; February, 1928.

(Quantitative results given of what effect changes in various condensers have on performance of filter circuit in B eliminators.)

R386 Wagner, K. Kettenleiter und Wellensiebe (filters). *Elektrot-Nachrichten Technik*, **5**, pp. 1–17; January, 1928.

(Theoretical application of filters to electrical communication.)

R386 McMeen, S. G. Filters. *Radio* (San Francisco), **10**, pp. 22–23; March, 1928.

(Explanation of filters. Chart given for the design of inductance coils with iron cores.)

R386 Replogle, D. E. Notes on the design of iron-core reactances which carry d.c. *QST*, **12**, pp. 23–27; Apr., 1928.

(Low pass type of filter for A and B battery substitutes described. Use of this type of filter for transmitting set plate supply.)

R387.1 Mason, R. B. The shielding efficiency of metals. *QST*, **12**, pp. 23–27; February, 1928.

(Measurements made of aluminum at radio frequencies show that value as shield can be predicted from its conductivity.)

R387.7 Walmsley, T. Notes on the design of radio insulators. PROC. I.R.E., **16**, pp. 361–72; March, 1928.

(Information on design of insulators used for radio transmission purposes.)

R388 Improved Duddell oscillograph outfit. *Jnl. Scientific Instruments* (London), **5**, pp. 103–107; March, 1928.

(Compact portable form of oscillograph. Applications in connection with electrical engineering practice.)

R388 Bedell, F. and Reich, H. J. The stabilized oscilloscope. *Radio Engineering*, **8**, pp. 17–20; April, 1928.

cathode-ray oscillograph with linear time axis—List of references included.)

R 400. Radio Communication Systems

R402 Grzybowski, W. W. Some investigations of short waves at Nijni-Novgorod. *QST*, 12, pp. 9–14; April, 1928.

(Description of apparatus and experimental work on high frequencies in the laboratory at Nijni-Novgorod, Russia.)

R412 Robinson, E. H. Some experiments with side band telephony on short waves. *Experimental Wireless* (London), 4, pp. 715–21; December, 1927.

(Advantages of side band transmission for frequencies above 1500 kc.)

R413 Peterson, E. and Keith, C. R. Grid current modulation. Bell System Technical *Jnl.*, 7, pp. 106–39; Jan., 1928.

(Application of grid current modulation in an experimental carrier telephone system. Tubes, filters and transformers discussed.)

R431 Armstrong, E. H. Methods of reducing the effect of atmospheric disturbances. Proc. I.R.E., 16, pp. 15–29; January, 1928.

(Balanced receiving system using difference between amplitude of strays and signal.)

R 500. Applications of Radio

R512 Gill, T. H. and Hecht, N. F. G. Rotating loop radio transmitters and their application to direction finding and navigation (Abstract of paper read before Wireless Section, Institute Elec. Engrs., London, on Jan. 4, 1928). *Experimental Wireless* (London), 5, pp. 85–88; Feb., 1928.

(Method of directional transmission developed whereby the direction finding equipment is eliminated on the aircraft and difficulties are reduced.)

R512 Smith-Rose, R. L. and Chapman, S. R. Some experiments on the application of the rotating beacon transmitter to marine navigation (Abstract of paper read before Wireless Section, Institute Elec. Engrs., London, on Jan. 4, 1928). *Experimental Wireless* (London), 5, pp. 88–90; Feb., 1928.

(Results of tests using rotating beacon transmitter for transmissions over land and sea. Beacon was calibrated.)

R512 Smith-Rose, R. L. A theoretical discussion of various possible aerial arrangements for rotating beacon transmitters (Abstract of paper read before Wireless Section, Institute Elec. Engrs., London, on Jan. 4, 1928). *Experimental Wireless* (London), 5, pp. 90–92; Feb., 1928.

(Three antenna arrangements for the rotating beacon transmitter are discussed theoretically. The uses of 3 systems of antennas, single coil, the double spaced coil, and the Adcock antenna for rotating beacons, are discussed.)

R520 Pratt, H. Radio guidance of aircraft. *Radio* (San Francisco), 10, pp. 19–20; Feb., 1928.

(Test flights to determine night and day variations in direction as given by crossed-coil beacon located in mountainous region.)

R520 Mitchell, W. G. W. London's new airport—Wireless the pilot of our airways. *Wireless World & Radio Review*, 22, pp. 130–32; Feb. 8, 1928.

(Description of new airdrome and receiving apparatus used for direction finding work.)

R520 Fassbender, H. Laboratorien und Forschungsarbeiten der Funkabteilung der Deutschen Versuchsanstalt fur Luftfahrt in Berlin—Adlershof (Laboratories and research work of Radio Division of Germany's experimental Laboratories for aeronautics in Berlin—Adler's field). *Zeitschrift für Hochfrequenztechnik,* **30,** pp. 173–176; Dec., 1927.

(Description of laboratories and radio apparatus used on airplanes in the test work carried out at Adler's field.)

R520 Linsmayer, R. Die drahtlose Einrichtung an Bord des A28 (Arrangement of wireless apparatus on board the plane A28). *Elektrotechnik und Massachinenbau* (Die Radiotechnik), pp. 23–26; March, 1928.

(Description of radio receiving and transmitting apparatus on the airplane.)

R520 Smith-Rose, R. L. Note on a special dial for time pieces to be used with rotating wireless or other beacons. *Jnl., Scientific Instruments* (London), **5,** pp. 93–96; March, 1928.

(Description of especially designed compass card dials for watches and chronographs used in connection with radio beacons.)

R520 Smith-Rose, R. L., and Chapman, S. R. Some experiments on the application of the rotating beacon transmitter to marine navigation. *Jnl.* I.E.E. (London), **66,** pp. 256–69; March, 1928.

(Results of test using rotating beacon transmitters for transmissions over land and sea. Beacon was calibrated.)

R520 Gill, T. H. and Hecht, E. S. Rotating loop radio transmitters and their application to direction finding and navigation. *Jnl.* I.E.E. (London), **66,** pp. 241–55; March, 1928.

(Method of directional transmission developed whereby the direction finding equipment is eliminated on the aircraft.)

R520 Smith-Rose, R. L. A theoretical discussion of various possible aerial arrangements for rotating beacon transmitters. *Jnl.* I. E. E. (London), **66,** pp. 270–79; March, 1928.

(The use of three systems of antennas; single coil, the double spaced coil and the Adcock antenna for rotating beacons is discussed.)

R531.2 Citizens' radio amateur call book. Published by Citizens' Radio Call Book Co., 508 So. Dearborn St., Chicago, Ill.

(Contains calls of commercial land and ship stations; amateur calls both foreign and domestic; special short wave stations—This company also issues separate book listing only broadcasting station calls).

R531.2 Short wave transmissions. *Radio Broadcast,* **13,** pp. 44–46; May, 1928.

(List of stations throughout the world working below 100 meters.)

R533 Tätz, P. Signalübertragung auf fahrende Züge mittels Wechselstrominduktion und Resonanz (signal transmission to moving trains by means of a.c. induction and resonance) *Telefunken Zeitung,* **11,** pp. 70–78; October, 1927.

(Description of apparatus used and experimental results.)

R533 Saglio, R. Receptions radiotelephoniques sur trains en marche (radio telephone reception on moving trains). *L'Onde Electrique,* **6,** pp. 589–602; Dec., 1927.

(Tests made by railroad company in France on reception of radio concerts from Eiffel Tower.)

R536 Eve, A. S. and Keys, D. A. Geophysical methods of prospecting. Bureau of Mines Technical Paper No. 420. Obtainable for 10 cents per copy from the Superintendent of Documents, Government Printing Office, Washington, D.C.

(A brief and elementary account of the principles involved including applications of radio. A short bibliography.)

R550 Dreher, C. Broadcast control operation. PROC. I.R.E., **16**, pp. 498–512; April, 1928.

(Considerations of audio-frequency elements of broadcast control system discussed. Methods of measuring energy levels, equalizing lines, etc. Coordinative and regulative functions of technical staff of broadcast system, precaution against breaks in program continuity discussed.)

R550 Eckersley, P. P. The design and distribution of wireless broadcasting stations for a national service. *Experimental Wireless* (London), **5**, pp. 189–97; April, 1928.

(The history of the development of British Broadcasting Co. stations is traced; experimental station at Daventry is described.)

R582 Schriver, O. Über Verstärkung in der Bildtelegraphie (On amplifiers for picture telegraphy). *Telefunken Zeitung*, **11**, pp. 78–84; Oct., 1927.

(Construction of amplifying apparatus.)

R582 Rowe, G. C. B. Television comes to the home. *Radio News*, **9**, pp. 1098–1100; April, 1928.

(Explanation of Alexanderson transmitter and description of television short wave receiver.)

R594 Fortschritte im elektrischen Nachrichtenwesen im Jahre 1927 im Deutschland (Progress in electrical communication in Germany during 1927). *Electrot-Nachrichten Technik*, **5**, pp. 33–39; January, 1928.

(Report of work along electrical lines including radio.)

R599 Deloraine, E. M. La station radiotelephonique de Prague (Radiotelephone station of Prague). *L'Onde Electrique*, **7**, pp. 21–32; January, 1928.

(Description of broadcasting station in Prague, Czechoslovakia.)

R 800. NON-RADIO SUBJECTS

510 Barclay, W. A. Dimensionality in wireless equations. *Experimental Wireless* (London), **5**, pp. 211–13; Apr., 1928.

(Points out practical use of the concept of dimensionality in checking the accuracy of equations developed in theoretical work, especially in the radio field.)

537.1 Howe, G. W. O. Some further problems in potential difference. *Experimental Wireless* (London), **5**, pp. 175–78; Apr., 1928.

(Treats some simple problems in the electric potential of dielectric rings. The questions which are considered are fundamental to a proper understanding of electric circuits.)

537.65 Cady, W. G. Bibliography on piezoelectricity. PROC. I.R.E., **16**, pp. 521–35; April, 1928.

(List of publications, patents and books on piezoelectricity up to beginning of 1928.)

621.313.7 Copper oxide rectifier. *Experimental Wireless* (London), **5**, pp. 1–2; January, 1928.

(Description and characteristic curve of rectifier.)

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED

June 6, 1928

Transferred to the Fellow grade

New Jersey	Boonton, Radio Frequency Laboratories	**Ballantine, Stuart**
	Boonton, Radio Frequency Laboratories	**Hull, Lewis M.**

Transferred to the Member grade

Illinois	Chicago, 7620 Phillips Avenue	**Falknor, Frank B.**
Kansas	Fort Leavenworth, Box 506	**De Remer, C. W.**
Maryland	Baltimore, 3100 St. Paul Street	**Wheeler, Harold A.**
Massachusetts	Cambridge, Massachusetts Institute of Technology	**Bowles, Edward L.**
	Winchester, Appalachian Road	**Browning, Glenn H.**
Michigan	Detroit, 504 Commerce Bldg.	**Brown, Jesse E.**
	Detroit, 504 Commerce Bldg.	**McNary, James C..**
New York	Little Neck, L. I., 4310 Bay View Avenue	**Di Blasi, John**
	Mount Vernon, 2 Park Avenue	**Houck, Harry W.**
	New York City, Radio Corporation of America, 233 Broadway	**Burchard, J. C.**
	New York City, 233 Broadway, Room 2001	**Adams, Ira J.**
	New York City, Radio Corporation of America, 66 Broad St.	**Finch, J.**
Wisconsin	Madison, University of Wisconsin, Sterling Hall	**Terry, Earle M.**

Elected to the Member grade

Connecticut	East Hartford, 400 Burnside Avenue	**Randall, J. Clayton**
Illinois	Chicago, Congress Hotel, c/o W. E. & M. Co.	**Evans, Walter C.**
New York	New York City., Bell Tel. Labs., 463 West Street	**Marrison, W. A.**
France	Paris, 16e, 58, Rue Michel Ange	**Cheftel, A. M.**

Elected to Associate grade

Arkansas	Little Rock, 1000 State Street	**Hofslatter, F. H.**
California	Los Angeles, 5157 Arlington Street	**Barlow, Roy G.**
	Los Angeles, 531 S. Mariposa Avenue	**Eldridge, Robert H.**
	Marshal, c/o Radio Corporation of America	**Wells, Wm. Bruce**
	Palo Alto, 219 Hawthorne Avenue	**Whittern, Ray**
	San Diego, 535 West University Avenue	**Conaughty, Robert B.**
	San Francisco, Pacific Gas & Electric Co., 245 Market St.	**Benson, F. S.**
	San Francisco, 202 Green Street	**Farnsworth, Philo F.**
	Sam Francisco, 742 Clement Street	**Maxwell, John F.**
	San Jose, 235 San Carlos Street	**Reynolds, F. T.**
Colorado	Arvada	**Larsen, Archie M.**
Connecticut	New London, 216 Pequot Avenue	**Helgren, Diva M.**
	Plainville, 9 Whiting Street	**Jones, H. Garfield**
Georgia	Tignall	**Talkington, Fletcher F.**
Illinois	Chicago, 205 North Crawford Avenue	**Kobberup, Jack R.**
	Chicago, 545 North Central Avenue	**Morey, W. J.**
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	Buffalo, 100 Taunton Place	Hobbs, Frank B.
	Buffalo, 236 Barton Street	Miller, George H.
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	New York City, 233 Broadway	Norton, Frederick A.
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	Schenectady, 21 Governor's Lane	Langdon, G. G.
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	Somerset, Bridgewater, Taunton Road	Ware, William
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India	Bangalore, Indian Institute of Science	Rangachari, T. S.
	Madras, Madras Club	Moone, Albert J. L.
New Zealand	Wellington, 39 Scarborough Terrace	Roberts, R. V.
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	Perth, 5th Avenue, Inglewood	Coxon, Walter E.

Elected to Junior Grade

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PROCEEDINGS OF

The Institute of Radio Engineers

Volume 16 August, 1928 Number 8

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GENERAL INFORMATION

The Proceedings of the Institute is published monthly and contains the papers and the discussions thereon as presented at meetings.

Payment of the annual dues by a member entitles him to one copy of each number of the Proceedings issued during the period of his membership.

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JONATHAN ZENNECK
of Munich, Germany
who comes to New York to receive the 1928 Institute Medal of Honor on September 5th.

JONATHAN ZENNECK

Jonathan Zenneck was born in Ruppertshofen, a small town in Wurttemberg, on April 15, 1871. After the usual education in the elementary schools, in the fall of 1885 he entered the Evangelical-Theological Seminary in Maulbronn, and in 1887 the seminary in Blaubeuren where he learned Latin, Greek, French, and Hebrew. In the fall of 1889 he enrolled in the Tuebingen University in the well-known Tuebingen Seminary where he studied mathematics and natural sciences. His teacher in physics was Professor Ferdinand Braun. In the spring of 1894 Professor Zenneck took the State examination in mathematics and natural sciences and immediately afterwards the examination for his doctor's degree.

During the summer of 1894 Professor Zenneck undertook zoological research at the Natural History Museum in London. From the fall of 1894–1895 he fulfilled the military service requirements in the First Naval Battalion (Marines) in Kiel, in which he later became a reserve officer. From 1895 to 1905 he was associated with the Physikalischen Institute in Strassburg, Alsace, first as assistant to Professor Braun and later as assistant lecturer.

Late in 1899 Professor Zenneck turned his attention to wireless telegraphy, conducting experiments along the lines indicated by Professor Braun. These experiments were mainly carried on on light ships in the North Sea. In the following year Professor Zenneck turned his attention to the many fundamental questions in wireless telegraphy which were unexplained and during the next year he turned to the theoretical and experimental explanation of the physical fundamentals of wireless telegraphy. The result of this work was the classical book "Electromagnetic Oscillations and Wireless Telegraphy," which appeared in 1906 and which, for so many years, was the standard textbook on the subject.

In the spring of 1905 he was appointed assistant professor at the Danzig Technical High School, and a year later became professor of experimental physics in the Braunschweig Technische Hochschule. In order to take part in experiments on the fixation of atmospheric nitrogen, in 1909 Professor Zenneck joined the staff of the Badische Anilin und Sodafabrik, one of the largest German chemical concerns.

In the fall of 1911 he returned to the Danzig Technische Hochschule as professor of experimental physics, and in 1913 he went to the Munich Technische Hochschule in the same capacity.

At the beginning of the World War Professor Zenneck went to the front as a Captain in the Marines. Early in December of 1914 he was sent to the United States as technical advisor for the Atlantic Communication Company, taking part in experiments with the machine senders in Sayville, and with patent processes.

After the United States entered the war he was interned first at Ellis Island and then in Fort Oglethorpe, Georgia, returning to Germany in July of 1919 to resume his duties as Professor of Experimental Physics at the Technische Hochschule in Munich. For the past two years he has been President of the Technische Hochschule.

Professor Zenneck has been a Fellow in the Institute since 1915.

THE INSTITUTE MEDAL OF HONOR

Each year the Institute awards a gold Institute Medal of Honor to that person who has made public the greatest advance in the science or art of radio communication, regardless of the time of performance or publication of the work on which the award is based.

The contribution may be an unpatented or patented invention which has been completely and adequately described in a scientific or engineering publication of recognized standing, and must be in actual operation.

It may also be a scientific analysis or explanation of hitherto unexplained phenomena of distinct importance to the radio art, although the application thereof need not necessarily be immediate.

The advance may further be a new system of traffic regulation or control; a new system of administration of radio companies or of service of steamship, railroad or other companies; a legislative program beneficial to the radio art, or any portion of the operating or regulating feature of the art.

In the past the medal has been awarded to the following engineers and scientists: Edwin H. Armstrong, 1918; E. F. W. Alexanderson, 1919; G. Marconi, 1920; R. A. Fessenden, 1921; Lee DeForest, 1922; John Stone Stone, 1923; M. I. Pupin, 1924; G. W. Pickard, 1926; L. W. Austin, 1927.

The Board of Direction of the Institute at its April 4th meeting decided that the 1928 Medal of Honor should be awarded to Professor Jonathan Zenneck for his contributions to original research on radio circuit performance and for this scientific and educational contributions to the literature of the pioneer radio art.

CONTRIBUTORS TO THIS ISSUE

Bower, Ward E.: Born in Schell City, Mo., April 4, 1884. Graduate Platts Commercial College, St. Joseph, Mo. Studied at Bush Temple of Music. Professional musician for many years. Studied mechanical electrical engineering, and mathematics by special instruction from Armour Institute, Chicago. Enrolled at University of Minnesota in radio engineering. Instructor in junior radio engineering at Wm. Hood Dunwoody Industrial Institute, Minneapolis, Minn. Designed and installed station WHDI and 9YX for the same institution. Radio manufacturing business in Minneapolis, Minn., 1920-1923. Since June 25, 1925, at Naval Research Laboratory as assistant radio engineer engaged in special research.

Crouse, George B.: Born at Arlington, N. J. in 1892; 1912–14 designer of high-frequency generators and automatic switching devices for the "Telharmonium"; 1914–21, department engineer in charge of gyrocompass and radio design with Sperry Gyroscope Company; 1921–22, chief engineer, Ford Instrument Company, working on gun fire control equipment; 1922–23, with Connecticut Instrument Company on development of acoustical apparatus; 1923 to date, development of power supply for radio apparatus with Rader Appliance Company and Conner Crouse Corporation.

Dellinger, J. H.: (See PROCEEDINGS for May, 1928).

Dreisback, Robert H.: Born at Fort Wayne, Indiana, March 12, 1907. Edison Lamp Works, Fort Wayne, Indiana, 1923. Test department, General Electric Company, Fort Wayne, 1926–27. Engaged in private work on short-wave communication, field intensity measurements, frequency measurement of short waves, broadcast station design, 1923–28. Received the A.B. degree in physics from Indiana University, 1928.

Goldsmith, Alfred N.: (See PROCEEDINGS for March, 1928).

Harris, Sylvan: Born at Philadelphia, Pa., May 5, 1898. Graduated from Central High School, Philadelphia; E.E. degree from University of Pennsylvania, 1922; engaged in power maintenance work, Midvale Steel Ordnance Company, Philadelphia, 1917–18; Leeds and Northrup, 1918; technical editor, *Lefax*, 1922–23; managing editor, *Radio News*, 1923–25; director of research and design, Stewart-Warner Speedometer Corporation, Chicago, 1926–27; 1927 to date in the laboratories of Brandes Products Corporation, Newark, N. J. Member in the Institute.

Hund, August: Born at Offenburg, Baden, Germany, December 17, 1887. Educated at Carlsruhe, Heidelberg and California; E.E. degree, 1911 and Dr. Eng. degree, 1913 from Carlsruhe. Research engineer in Research Laboratory of the General Electric Company at Schenectady, 1912–14, part of the time in Dr. Steinmetz's private laboratory; assistant professor of physics and electrical engineering, University of Southern California, 1915–17; consulting research engineer in San Francisco, 1918-22. Since November, 1922, electrical engineer at the Bureau of Standards doing research work in radio and electro-acoustics. Fellow in the Institute.

Kimmell, William J.: Born at Pittsburgh, Pa., May 4, 1905. Undergraduate work in physics at Carnegie Institute of Technology and degree of B.S. in physics, 1926. Since employed as research engineer in vacuum tubes and vacuum-tube circuits by the Westinghouse Electric and Manufacturing Company at East Pittsburgh. Also has taken graduate courses in physics at Harvard University, University of Chicago, and University of Pittsburgh.

Ramsey, R. R.: Born at Morning Sun, Ohio, April 11, 1872. Received A.B. degree from Indiana University, 1895; A.M. degree, Indiana University, 1898; Ph.D. degree, Cornell University, 1901. Instructor, University of Missouri, 1901–03; assistant professor, associate professor and professor, Indiana University, 1903 to date. Author of a number of papers relating to electrolytic conduction, radio-activity and radio telegraphy. Author of "Experimental Radio." Fellow of American Physical Society and Member of the Institute of Radio Engineers.

Thomson, J. M.: Graduated from the University of Toronto in applied science, 1924. General Electric test course, 1924, specializing in radio receivers; transformer engineer, English Electric, 1925–26; 1927 to date in charge of Toronto Research Laboratory of Ferranti Electric, Limited.

van der Pol, Balth.: Born at Utrecht, Holland, January 27, 1889. Received doctor's degree from University of Utrecht, 1916. Research work in radiotelegraphy and allied subjects at Cambridge University, England, 1917. Conservator of Physical Research Laboratory, Teyler's Institute, Holland, 1919. Research Laboratory, Philips' Glowlamp Works, Ltd., Eindhoven, Holland, 1928. Member of the Institute.

Wheeler, L. P.: Born at Bridgeport, Conn., July 27, 1874. Received the Ph.B. degree from Yale University, 1894, and the Ph.D. degree in 1902; laboratory assistant in physics, Sheffield Scientific School, Yale University, 1895–97; assistant, 1897–01; instructor, 1901–06; assistant professor, 1906–23; and associate professor, 1923–26. Physicist, U. S. Naval Research Laboratory, 1926 to date.

Wright, James Warren: Born at Springfield, Ohio, 1899. Received A.B. degree from Ohio Wesleyan and the M.A. degree from Ohio State University, with additional graduate work at Ohio State University. Teaching experience in physics department of Ohio State, Ohio Wesleyan, and Syracuse Universities. Member of technical staff, radio division, U. S. Naval Research Laboratory since June, 1926. Work has been in connection with precision, electrical and radio measurements, piezoelectric crystal and vacuum-tube transmitter circuits.

INSTITUTE ACTIVITIES

Dinner to Professor Zenneck

Members of the Institute are afforded the very unusual opportunity of honoring a distinguished fellow-member from a foreign country who comes to this country to receive the Institute Medal of Honor and to address the Institute. Professor Jonathan Zenneck arrives in the United States early in September to be the guest of honor at a dinner which the Institute gives him at the Hotel McAlpin, Broadway and 34th Streets, New York, on the evening of September 5th.

Following the dinner, a meeting of the Institute will be held in the Engineering Societies Building, 33 West 39th Street, at which Professor Zenneck will be presented with the Medal of Honor and will deliver a paper on, "What Science Owes to Radio Telegraphy."

All members and their friends are invited to attend the dinner, which will be informal. Admission to the dinner will be by ticket only. Reservations, at $3.00 per plate, must be accompanied by remittances, and *must* be in the hands of the Secretary of the Institute not later than August 23rd. Members' wives and guests are invited to the dinner and the Institute meeting which follows.

Preliminary 1928 Standardization Reports

Only a few copies of the preliminary draft of report of the Institute's Committee on Standardization for 1928 are now available. All persons interested in radio standardization should receive a copy of the preliminary draft. Copies may be obtained free of charge by addressing a request to the Institute office.

Publication Program

Due to the greatly increased number of pages which are being published in the Proceedings each month, the publication program for the Proceedings has been considerably expanded. Manuscripts submitted for publication can appear in the Proceedings in approximately one quarter of the time formerly required.

The Institute's Committee on Meetings and Papers will welcome any manuscripts on any phase of radio or its closely allied arts. Manuscripts should be forwarded to the office of the Institute, 33 West 39th Street, New York City.

NEW MEMBERS TO BE ELECTED

On page 1156 of this issue will be found a list of applicants for the several grades of membership. All of these have been favorably acted upon by the Committee on Admissions for submission to the Board of Direction for election at its September 5th meeting. Members objecting to the transfer or election of any persons listed should communicate with the Secretary of the Institute on or before September 2nd, 1928.

Institute Meetings

CLEVELAND SECTION

The first dinner-meeting of the Cleveland Section was held in the Hotel Winston on Friday evening, June 8th. The meeting was one of the most enjoyable and profitable ones of the year.

Donald McNicol, Chairman of the Institute's Committee on Sections, was present and delivered an address. Musical entertainment was provided during the dinner. The first speaker was Edward L. Viets, who gave a discussion of present day radio conditions from a strictly humorous point of view.

A comedy skit in black face worthy of "big time" was presented by Messrs. Schonvisner and Hirshfield, senior students at Case School of Applied Science.

Chairman John R. Martin in introducing Mr. McNicol spoke of the excellent co-operation and friendly assistance which is afforded by the New York Headquarters. Informal discussion followed Mr. McNicol's talk in which Section problems were discussed.

The attendance at this meeting was forty-seven.

DETROIT SECTION

On April 20th a meeting of the Detroit Section was held in the Detroit News Building. E. D. Glatzel, chairman of the Section, presided.

A. B. Buchanan, of the Detroit Edison Company, presented a paper on "Sources and Mitigation of Radio Noises" in which the routine employed by the larger electric power utilities in handling complaints of radio disturbances was explained, special reference being made to the problems presented by the popularity of the new a.c. receiver. The causes of trouble with these receivers, in some cases, were explained and remedies suggested. The fallacy of the "leaky transformer" was pointed out. A number of the more common peculiarities of radio disturbances were described and it was pointed out that experience is the most valuable asset in locating sources of radio interference, although difficult cases of trouble are frequently located as a result of persistence rather than brilliance on the part of the investigator.

A device for recording interference was described. This device is designed for use in cases of intermittent troubles at some distance from headquarters.

An analysis of more than 6,000 cases of trouble reported to the power company showed a surprisingly small number of cases of trouble actually caused by their equipment.

Thirty-five members and guests attended this meeting.

The Detroit Section held a meeting on May 18th in the Detroit News Building. Earle R. Glatzel presided.

The speaker, W. W. Brown of the radio engineering department of the General Electric Company, presented a paper on "Directive Antennas." The paper contained a detailed explanation of the fundamental principles of directive radiation from a simple arrangement of two radiators. The effect of time and space relations was considered. Patterns showing directive properties of single unit radiators were presented, including reference to horizontal and inclined directivity. Descriptions of various types of directive antennas in commercial use were given. These included the extended half wave, parabola, and the degenerate parabola with infinite and zero focal adjustments. Theoretical improvements by use of certain of these types compared with single units were given. An outline of additional data which is being obtained, and lines of further scientific investigation which are being followed were presented.

San Francisco Section

The June meeting of the San Francisco Section was held in the Engineers' Club on June 20th. S. W. Edwards, radio supervisor of the 8th Inspection District, and J. E. Brown, radio inspector from that district, were present as guests, and spoke briefly upon the radio test cars of the Department of Commerce and their use in field-strength measurements. Elmer L. Brown, service engineer of the California Victor Company, presented a paper, "A Unique Portable Instrument for Radio Testing." He outlined, briefly, the service requirements leading up to the design of the instrument, and described in detail the circuits and construction which permit all necessary tests on a radio receiver to be made with a single piece of equipment, readily carried and easily used.

The attendance at this meeting was thirty-two.

Committee Work

A 1928 Committee on Nominations, with membership as follows, has been appointed by President Goldsmith: J. H. Dellinger, Chairman; Donald McNicol, R. H. Marriott, Arthur Batcheller, and L. M. Hull.

This committee is to make recommendations to the Board of Direction as to possible candidates for the offices of President, Vice President, and members of the Board of Direction (two) at the September 5th meeting of the Board. The Committee desires to receive suggestions from the membership as to the names of possible candidates for these offices.

As authorized in the amendment to the Constitution of the Institute, dated January, 1921, "Nomination by Petition shall be made by letter, addressed to the Board of Direction, setting forth the name of the proposed candidate and the office for which it is desired he be nominated. For acceptance, a letter of Petition must reach the Board of Direction on or before October 15th of any year, and shall be signed by at least thirty-five Fellows, Members, or Associates."

COMMITTEE ON ADMISSIONS

A meeting of the Committee on Admissions was held in the office of the Institute on July 12th. Messrs. F. H. Kroger, E. R. Shute, and H. F. Dart were present.

The committee passed upon eighteen applications for transfer or election to the Member or Fellow grade.

Erratum

In connection with the paper by W. A. Marrison appearing in the July, 1928 issue of the PROCEEDINGS, the following caption was omitted from Figure 3, page 979:

Fig. 3. Illustration of the Effect of Using a Thermal Attenuating Layer.

Material of Attenuating Layer	Temperature Variation at B	
	For 1 Min. Period	For 10 Min. Period
Wool felt	±0.002	±0.14 deg.
Equal alternate layers of felt and copper	±0.0002	±0.06 "
Felt and copper in tapered arrangement as described in text.	±0.00006	±0.03 "

A NEW TYPE OF STANDARD FREQUENCY PIEZO-ELECTRIC OSCILLATOR*

By

Lynde P. Wheeler and Ward E. Bower

(Naval Research Laboratory, Bellevue, Anacostia, D. C.)

Summary—In this paper is described a system for producing alternating current of audible frequency of a very high degree of precision. The chief novelty in the system consists in the manner in which the energy necessary to sustain the oscillations of the quartz bar is returned to it. This "feed-back" is accomplished acoustically and the extremely loose coupling thereby secured insures that the generated frequency is practically uninfluenced by inertia effects of electrodes or variations in tube constants.

There is also described the installation of the auxiliary apparatus to permit the continuous operation of the oscillator over very long periods of time and to measure the generated frequency with the highest possible precision.

SEVERAL methods of using a crystal-controlled vacuum-tube oscillator as a standard of frequency are described[1] in the literature. In none of these methods is the generated frequency exactly that of the free mechanical oscillations of the quartz plate. This is because the combination of piezo-electric quartz plate and the vacuum tube with its associated circuits form a coupled system of more than one degree of freedom, and moreover one in which the coupling has to be fairly "tight" in order to secure the energy necessary for self-sustaining operation. From the point of view of its use as a frequency standard, the fact that the generated frequency is not the natural frequency of the quartz plate would make no difference, provided that the former remained constant. Unfortunately, however, it is precisely those factors which cause the generated to differ from the quartz-plate frequency that are subject to variations difficult to control where the highest precision is desired. The most important of these factors is the internal capacity between the grid and plate of the tube. Its magnitude affects the numerical value of the generated frequency in two ways. First, being a condenser

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* Original Manuscript Received by the Institute, June 20, 1928. Read at April, 1928 meeting of the International Union of Scientific Radiotelegraphy.

[1] Cady, Proc. I. R. E., **10**, p. 83; April, 1922.
Pierce, *Proc.* Am. Acad. Arts and Sciences, **59**, p. 83; Oct., 1923.
Dye, *Proc.* Phys. Soc. London, **38**, p. 399; Aug., 1926.
Hund, Proc. I. R. E., **14**, p. 447; Aug., 1926.
Horton and Marrison, Proc. I. R. E., **16**, p. 137; Feb., 1928.
Worrall and Owens, Proc. I. R. E., **16**, p. 778; June, 1928.

in series with the capacities representing the quartz plate in its electrical analogue,[2] it affects the equivalent LC value; and second, forming as it does the coupling element of the system it affects the frequency through the dependence of the latter on the coupling coefficient. Thus, any variation of this tube capacity with time is doubly important, necessitating rather frequent calibrations, and in case of tube replacements it may require readjustment of circuit constants.

The very considerable experience with this type of oscillator obtained at the U. S. Naval Research Laboratory leads us to believe that these inherent limitations in the constancy of the oscillator represent, under ordinary laboratory conditions, an unavoidable error of somewhat less than 0.001 per cent. Under the *best* conditions—that is, at the time of a calibration and with better than ordinary temperature control—it is possible to be certain of the oscillator frequency values to slightly less than 0.0001 per cent. Such accuracy is, however, not easily obtainable as a routine procedure.

Now, while measurement of frequency to better than one part in 100,000 is amply accurate for most purposes in the radio art today (this means an uncertainty of less than 200 cycles in 20 megacycles), it would be rash to say that it will suffice even five years hence. Further, it does not seem practicable at present to attain even this precision of frequency in the radiations of service transmitters, where 0.01 per cent would seem to represent the best modern practice. As the same principles are involved in the design of crystal-controlled transmitters as in laboratory apparatus for the measurement of frequency, it is reasonable to expect that improvements in the latter will be applicable to the former, thus raising transmitter performance to a point (in respect to frequency stability) more nearly adequate to the demands of the radio art of the present, not to mention those of the future. Finally, if any further justification for playing with the ultimate decimal places is needed, it must be urged that mere intellectual curiosity to explore the limits of the possibilities of such a marvelous device as the piezo-electric oscillator is the best possible reason for doing so.

In casting about for means of improving the performance of the usual types of standard oscillator, it seemed logical to look for improvement by reducing the coupling between the mechanical and electrical features of the device. This would entail, of course, a reduction in the output energy available, but that, from the point of view of a measuring apparatus, was a secondary consideration. In order to attack intelligently the problem of reducing the coupling it seemed advisable to obtain more information than is available in the literature as to the magnitude of the voltages obtainable from the mechanically vibrating quartz plates. In other words, since reducing the coupling would mean a reduction in the energy transferred between what may be regarded as the primary and secondary of a transformer, itis necessary to have some idea as to the amount of energy available in the primary, in order to determine the feasible limit to which the coupling may be reduced.

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[2] See, e.g., Dye, loc. cit.

For this purpose an apparatus was constructed for mechanically vibrating a quartz plate at frequencies and amplitudes readily controllable, and the piezo-electric voltages developed were measured with a sensitive vacuum-tube voltmeter. It is unnecessary at this time to go further into the details of this investigation, as owing to various causes it is as yet incomplete. Suffice it to say that it soon developed that even with very small amplitudes and at frequencies far removed from those proper to any of its known modes of oscillation, there were piezo-electric differences of potential produced of the order of 100 microvolts. It then became immediately obvious that given sufficient amplification, the driver of our mechanical vibrator could be actuated, and we would have a self-sustained oscillator, in which the vacuum tubes would play the part of amplifiers only and not function at all as a coupling or regenerative element between the mechanical and electrical oscillations. Thus, variations in the interelectrode capacities or in the external circuits due to any cause whatever could not affect the frequency except as slight changes in the amount of amplifier output energy furnished the driver of the mechanical vibrator might change the amplitude of its output and hence conceivably, by varying the amplitude of the quartz-plate oscillations, alter the frequency of the latter. As a little consideration showed that such changes could with careful design be reduced to insignificance, there remained only the problem of the method to be used to vibrate the quartz mechanically.

There are, of course, a number of more or less obvious solutions to this problem, most of which, however, involve rather formidable difficulties in the way of complicating the apparatus, or make the application of the necessary temperature control awkward. The simplest solution that suggested itself, and that finally adopted, was to vibrate the quartz plate by a resonant air column. This naturally restricts the frequency obtainable to relatively low values, but otherwise adapts itself so perfectly to the requirements of a standard of the highest precision, that there was no question of its adoption after the idea had presented itself.[3] In all the changes of form which the apparatus had undergone up to and including that which it is the purpose of this paper to describe, this feature of the acoustical "feed back" of the energy necessary to sustain the oscillations has remained unchanged. In all the stages of development the amplified piezo-electric voltage operates a loudspeaker element, which in turn actuates a tunable air column and resonator from which the quartz bar receives its sustaining energy. All of the essential changes which have been made are in the method of

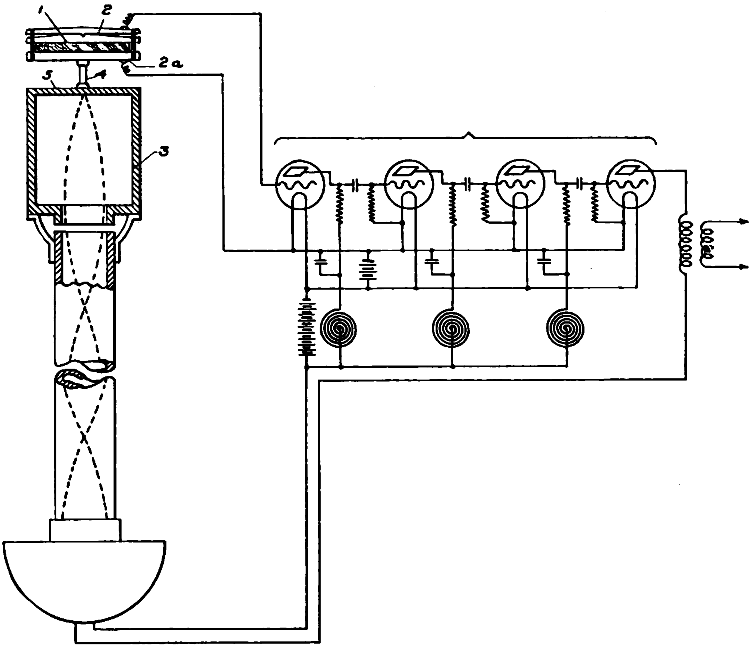

Fig. 1

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[3] The idea of maintaining the oscillations of the bar by the return of energy to it acoustically was first suggested by Mr. Bower.

mounting the bar and in the manner of collecting the piezoelectric charges which its motion develops. It will be profitable to trace in some detail certain of these changes because one point that emerged is of considerable importance in its bearing on the function of the apparatus as a primary frequency standard.

The first form which the apparatus took is shown schematically in Figs. 1 and 1a. With what has been said above the functioning of the various parts will be evident without further description. Attention is here directed only to the method of mounting of the quartz bar (1), which is so cut that the electric axis is parallel to its shortest and the optic axis to its intermediate dimension. It will be noted that the bar is clamped at the center between small electrodes (2) and (2a) insulated from each other and connected to the amplifier. It receives its energy from the resonator (3) through the stem (4) rigidly attached to the thin-walled bottom (5) of the resonator. The electrodes were attached to a place of maximum stress in order to get as large a voltage as possible and so make extreme amplification unnecessary.

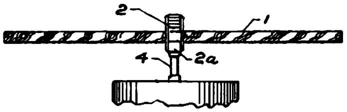

Fig. 1a

Although this form of the apparatus functioned well it was feared that the existence of a clamp with a member of its assembly inferior both mechanically and as an electrical insulator to the material of the bar would lead to a damping coefficient for the oscillations of the latter which would be variable with time and thus cause variations in frequency of the order of magnitude of those inherent in the usual type of standard oscillator. It was in addition desired to leave the bar free of electrodes if possible in order to get away from any complications due to loading effects of attached metal. Hence the mounting shown in Fig. 2 was substituted, the rest of the apparatus remaining as in Fig. 1.

In this form of mounting the bar (1) was supported on cork or felt strips (2) and (2a) placed at the nodes of the mode of vibration which it was desired to use. The piezo-electric voltage was developed on auxiliary crystal plates (3) and (4) which were vibrated by their attachment to the walls of the resonator (2).

With this arrangement the bar does not need to be of piezoelectric material, and in fact the apparatus has been used successfully with metallic bars. This method of mounting represented a material advance over the previous one in that the bar is left practically free to execute its own transverse vibrations as a bar supported at the nodes uninfluenced by mechanical reactions of electrodes. However, an electric reaction which might be unfavorable still remained, as was shown by the following observation. When the bar was of crystalline quartz and the air column was adjusted to return the energy in the proper phase to sustain the oscillations, it was noticed that the bar would not vibrate continuously if it were turned over so that its other face was opposite the resonator opening. This behavior is due to the fact that the bar is urged to move not only by the alternate rarefactions and condensations of the air at the mouth of the resonator, but also by the electrostatic attractions and repulsions between the charge developed on the near surface of the bar and that on the electrode (5) which originates from the motion of the auxiliary crystal plates (3) and (4). If the two resulting forces conspire the oscillations of the bar will persist, but if they oppose the oscillations quickly die out.[4]

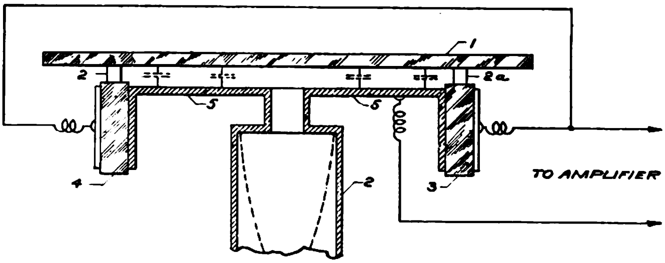

Fig. 2

Now this effect, though interesting, is not of particular importance, except for the rather surprising magnitude of the electrostatic forces involved. This immediately suggested that the auxiliary crystals as well as electrodes on the bar could be dispensed with and resulted in the finally adopted form of mounting shown in Fig. 3. In this mounting the bar is held at the nodes of its fundamental mode (very closely 1000 cycles) by stout silk threads between spring metal supports. The electrodes for various reasons connected with questions of design are placed much nearer the bar than is necessary for successful operation. A bar has been satisfactorily operated with electrodes distant as far as five or six centimeters.

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[4] A detailed study of the type of vibration utilized in this oscillator is in course of preparation.

We thus arrived at a design for an oscillator in which, (1) the coupling between the mechanical and electrical oscillations involved has been very much reduced; (2) the values of the tube constants (provided they are such as to give sufficient amplification) cannot affect the generated frequency except through the effect of very small changes in the amplitude of the oscillations of the bar due to changes in the energy supplied by the loudspeaker element to the air column; (3) inertia effects of electrode loading have been reduced to an unavoidable minimum (in the absence of the possibility of utilizing a vacuum mounting); (4) the generated frequency is more nearly that determined by the dimensions and elastic properties of the quartz plate than has previously been possible.

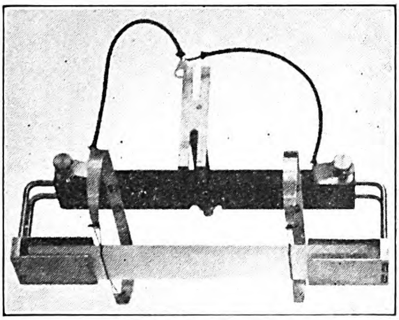

Fig. 3

In order to test out quantitatively the performance of this new system a rather elaborate equipment has been designed and installed at the U. S. Naval Research Laboratory.[5] The mounting of the quartz bar together with the amplifier and the acoustic feed-back system, the latter provided with micrometer tuning adjustment, have been combined into a self-contained unit (see Fig. 4), which when in service is immersed up to the point (a) in the water bath of a large thermostat. The latter is located in a triple-walled booth provided with automatic temperature control. Both the booth and the water thermostat heaters and stirrers are operated from the service mains with automatic switching devices to provide storage-battery supply in case of failure of the regular service. The amplifier is provided with two sets of tubes, two filament and two plate supply batteries. By means of a control box located on the outer wall of the booth the change from one set of tubes to the other, the change from one set of batteries to the other, and the charging of the set of batteries not in use, is provided for without interrupting the oscillator. These precautions are taken to ensure continuous operation at a definite temperature over long periods of time.

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[5] For the details of this design Mr. Bower is responsible.

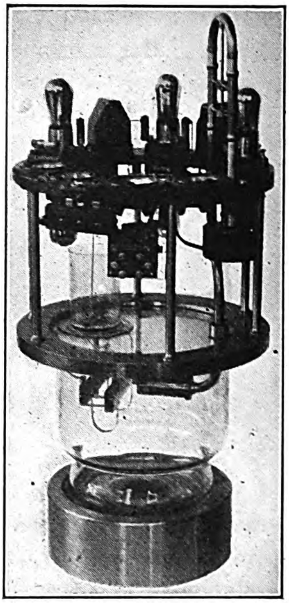

Fig. 4

The equipment provided for determining the value of the generated frequency consists of four units: (1) an amplifier fed from the output of the oscillator unit which in turn supplies the current to operate; (2) a synchronous motor clock; (3) a radio receiver unit for the reception of the U. S. Naval Observatory time signals from Arlington and which, when those signals arrive, automatically starts; (4) a tape recorder on which they are printed simultaneously with signals from the synchronous motor clock which latter are sent out ten times in each of its seconds. The amplifier driving the synchronous motor clock like that in the oscillator unit is provided with two sets of tubes and is operated from the same batteries through the same control box. The clock gearing is such that if supplied with exactly one-thousand-cycle current it will keep correct time. Thus, at each period that the time signals are sent out, we obtain a permanent record from any two points of which the frequency generated by the oscillator can be determined with very considerable accuracy, which can be made (by choosing a long enough time interval) to be as great as that of the time signals themselves. That is, of course, providing the oscillations of the quartz bar remain constant in frequency throughout the whole interval. In any case, the measurements made as outlined above will give the *average* value of the frequency of the bar to that accuracy.

To check the constancy of frequency over such long periods of time, a second oscillator unit has been provided equipped with a duplicate set of controls and a second clock, printing on the same tape recorder. The two oscillator units are immersed in the same thermostatically-controlled bath. Switching arrangements are provided so that either clock may be recorded against the time signals, or one clock against the other, or either clock against any other standard frequency source. Thus, from these permanent daily records of the performance of the two bars it would seem that in time we should obtain very exact information as to the degree of constancy of the oscillations of the bars.

At the present time the equipment has been in operation for too short a time to permit a quantitative statement of its performance to be made. In fact, the second oscillator unit and the tape recorder have but just been installed. However, from such tests of the first oscillator as have been made without the recorder, it is fairly certain that the *average* frequency of its quartz bar has not varied by as much as one part in a million and that in all probability the variations are much smaller than that.

Before the final measurements on the relative rates of the clocks with respect to each other and of each one separately with respect to Naval Observatory time can be made, two experiments will have to be performed. First, the effect on the generated frequency of slight changes in the amplitude of oscillation of the bars must be determined. This can be conveniently accomplished with the aid of the micrometer adjustment provided for the air column of the acoustic feed-back system.

Secondly, the temperature coefficient of frequency for this type of vibration must be determined. This will be accomplished, of course, by varying the temperature at which the thermostat operates.

From one's instinctive confidence in the immutability of quartz, aged as it has been through geologic time, as well as from our experimental experience with short time comparisons of the relative frequencies of the older type of crystal-controlled oscillators, and from that which we have already had with the present apparatus, it seems reasonable to expect that our measurements will show when extended over a period of some years that the ratio of bar number one's second to bar number two's second will remain constant to a higher degree of accuracy than the ratio of either's second to that of the Naval Observatory's second. There is, however, a consideration which renders the prediction of such a result uncertain. It may be argued that although crystalline quartz as found in nature is a very permanent substance, it nevertheless has not been subjected to the particular kind of periodic stresses which it must now suffer, and that in consequence there may occur slight secular changes difficult or impossible to determine.

With the idea of accentuating any possible changes of this nature, one of our bars is polished and the other left relatively rough ground. Thus, as in the case of unaged metals after machining operations, we should be able to detect any changes of this nature more readily. But "sufficient unto the day is the evil thereof"! It will be time to face the consequences of the advent of a more constant timekeeper than the earth, if and when such measurements as we expect to make in the next few years show that it has really arrived.

THE EFFECT OF REGENERATION ON THE RECEIVED SIGNAL STRENGTH*

BY

BALTH. VAN DER POL

(Physical Laboratory, Philips' Glowlamp Works, Ltd., Eindhoven, Holland)

Summary—It is the purpose of this paper to give a theory of the effect of regeneration using the solution of a non-linear differential equation, and to present experimental verification of the theory.

It is shown that: (*a*) *as a first approximation, detection has no effect on the radio-frequency grid voltage developed under the influence of an incoming signal,* (*b*) *the amplification obtained through regeneration equals the two-thirds power of the ratio of the "grid space" to the amplitude obtained with zero regeneration. It is apparent from* (*b*) *that much greater gain is obtained through regeneration with weak signals than with strong signals.*

The verification of the theory is made with a circuit arrangement operating at 500 cycles per second. Application is made to radio frequencies using this as a model of a high-frequency system, following a theorem for model systems which is stated.

THE considerable increase in signal strength obtainable through the use of regeneration is well-known. It is also common knowledge that this increase in signal strength is considerably greater when the incoming signal is weak than when it is strong. It is the purpose of this paper to provide a non-linear theory of this effect of regeneration and the experimental verification thereof.

Suppose a triode system of one degree of freedom, as in Fig. 1 where L, C, r form the tuned circuit and the mutual induction M provides the regeneration.

Let further an emf, $E \sin \omega_1 t$, representing the "incoming signal," be applied to the oscillatory system. Neglecting the grid current and calling the current in the LCr circuit i, and the deviation of the anode current from its steady value i_a and the alternating grid P.D. V_g, we have

$$L\frac{di}{dt}+ri+\frac{1}{C}\int idt-M\frac{di_a}{dt}=E\sin\omega_1 t \tag{1}$$

$$\frac{1}{C}\int idt=V_g \tag{2}$$

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* Original Manuscript Received by the Institute, May 26, 1928. Paper read at the meeting of the International Union of Scientific Radiotelegraphy, October, 1927.

In the "tuned grid" circuit, the variations of anode potential are usually small compared with the variations of the grid potential, so that as a first approximation, the anode current variation i_a is a function of the grid potential variation only, i.e.,

$$i_a = \int (v_g) \tag{3}$$

In order to avoid complexity we neglect the grid current and we therefore imagine a negative grid bias to be provided. Our system of coördinates in the i_a, v_g plane is therefore as shown in Fig. 2. This differs from the usual notation insofar as the zero point of the coördinate system is shifted towards the steady d.c. position round which the oscillations occur.

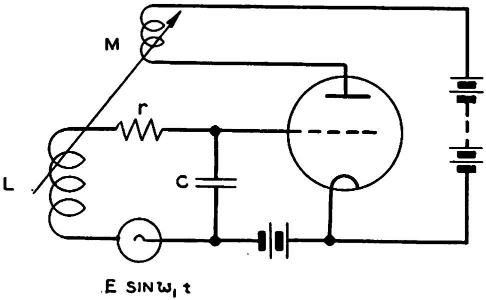

Fig. 1

We approximate equation (3), representing the curved plate current-grid voltage characteristic of the triode, by the cubic:

$$i_a = S_1 v_g + S_2 v_g^2 - S_3 v_g^3 \tag{3^a}$$

where S_1 is the usual "mutual conductance" for infinitesimal grid-potential variations. S_2 and S_3 are further determined by the form of the characteristic. As the latter bends round both at the top and at the bottom we write in (3a) $-S_3 v_g^3$ instead of $+S_3 v_g^3$.

The elimination of i and i_a from (1), (2), and (3a) results in:

$$\frac{d^2 v_g}{dt^2} + \left\{ \left(\frac{r}{L} - \frac{MS_1}{LC} \right) - \frac{2MS_2}{LC} v_g + \frac{3MS_3}{LC} v_g^2 \right\} \frac{dv_g}{dt} + \omega_0^2 v_g = \omega_0^2 E \sin \omega_1 t. \tag{4}$$

where

$$\omega_0^2 = \frac{1}{LC}$$

Calling further

$$\frac{r}{L}-\frac{MS_1}{LC}=\alpha,$$

$$\frac{MS_2}{LC}=\beta,$$

$$\frac{MS_3}{LC}=\gamma,$$

(4) becomes

$$\ddot{v}_g+(-\alpha+2\beta v_g+3\gamma v_g^2)\dot{v}_g+\omega_0^2 v_g=\omega_0^2 E \sin \omega_1 t. \qquad (4^a)$$

which is a non-linear inhomogeneous differential equation of the second order with non-linear resistance terms. The usual elementary approximation of a linear characteristic would make $\beta=\gamma=0$ and for resonance (i.e. $\omega_0^2=\omega_1^2$) and critical regeneration (i.e. $\alpha=0$) the developing grid voltage V_g would become infinite. In order to obtain a satisfactory theory of the response of a regenerative triode system to an empressed emf a non-linear problem must therefore be solved.

The equation (4a) was fully considered in a former paper[1]; from the results obtained there it follows that the steady state solution in the neighborhood of resonance ($|\omega_0-\omega_1|\ll\omega_1$) can be written:

$$v_g=C \sin (\omega_1 t+\phi)$$

where the amplitude C of the resulting grid potential variation is given by:

$$C^2\left\{4(\omega_0-\omega_1)^2+(\alpha-\tfrac{3}{4}\gamma C^2)^2\right\}=\omega_0^2 E^2 \qquad (5)$$

a cubic equation in C^2. It is further seen from (5) that the asymmetrical term in (4a) (with β) (which determines the detection) has, in the first approximation here considered, no influence on the resulting V_g. *Therefore as a first approximation detection has no effect on the r.f. grid potential difference developing under the influence of an incoming "signal."*

Further if (5) is compared with the usual linear case as represented by

$$C^2\left\{4(\omega_0-\omega_1)^2+\alpha_0^2\right\}=\omega_0^2 E^2,$$

it follows that in the non-linear case $(\alpha-\frac{3}{4}\gamma C^2)$ is substituted for

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[1] *Phil. Mag.* **3,** 65 1927.

α_0, i.e. the system behaves as if for the resistance r a new resistance r' were substituted of a value given by

$$r' = r - \frac{M}{C}\left(S_1 - \frac{3}{4}C^2 S_3\right) \tag{6}$$

which therefore depends upon the amplitude C already present in the system.[2]

When the regeneration is pushed so far that

$$MS_1 > rC \tag{7}$$

it is seen from (6) that the first order differential resistance of the system becomes negative and therefore the system has the tendency to oscillate spontaneously. However, as was shown in the *Phil. Mag.* paper quoted above, the forced oscillations may suppress the development of the free oscillations. The phenomenon manifests itself through the presence of a "silent region" extending at both sides of resonance. As was shown by Professor Appleton[3] the width of this silent region is determined by the amplitude of the incoming signal. For strong signals this width is given by

$$\frac{\omega_0 - \omega_1}{\omega_0} = \pm E\sqrt{\frac{3\gamma}{\alpha}} \tag{8}$$

It is of interest to investigate the resultant grid amplitude b if (a), the system is tuned exactly to resonance, i.e.,

$$\omega_0 = \omega_1$$

and (b), it is brought on the verge of free oscillation, i.e., when

$$S_1 M = Cr. \tag{9}$$

For this case we at once obtain from (5)

$$\tfrac{3}{4}\gamma C^3 = \pm \omega_0 E, \tag{10}$$

which expression will now be considered in detail.

First, it follows from (10) that *the resulting grid amplitude is proportional to the cube root of the emf applied to the system.*

Further, it is easy to assign an approximate value to γ directly from the triode characteristic. Referring to (3a), (4a) and Fig. 2, and taking the symmetrical case for which $S_2=0$, it follows that a good approximation to S_3 can be found from the maximum and minimum value of the cubic (3a), for $i_a=i_{a\ \max}$ when

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[2] This property of the non-linear system was first derived in 1920. See Balth. van der Pol, *Radio Review*, Nov., Dec. 1920.

[3] E. V. Appleton, *Proc.* Camb. Phil. Soc., **23**, 231, 1923.

$$v_g^2=\frac{S_1}{3S_3}.$$

Calling, therefore, the grid voltage change necessary to bring the anode current from zero to its saturation value the "grid space" and designing it by V_{g0} (see Fig. 2), we obtain

$$V_{g0}^2=\frac{4S_1}{3S_3} \tag{11}$$

hence

$$\gamma=\frac{M}{LC}\cdot\frac{4S_1}{3V_{g0}^2}.$$

But, when the system has critical regeneration (9) obtains, i.e.,

$$M=\frac{Cr}{S_1}$$

hence

$$\gamma=\frac{r}{L}\cdot\frac{4}{3V_{g0}^2}.$$

When we further call V_{g1} the grid-voltage amplitude which would be obtained with no regeneration at all, ($M=0$), or (which is the same thing) with reduced filament current, we find from (5)

$$v_{g1}^2=\frac{\omega_0^2L^2}{r^2}\cdot E^2,$$

hence we obtain from (10):

$$C^3=v_{g1}\cdot V_{g0}^2$$

or

$$C=\sqrt[3]{v_{g1}\cdot V_{g0}^2} \tag{12}$$

reading in words:

The grid amplitude developing in resonance and with critical regeneration equals the cube root of the product of the grid amplitude *which would be obtained with no regeneration at all, and the square of the "grid space"*, as defined by Fig. 2.

(12) can further be written:

$$\frac{C}{v_{g1}}=\left(\frac{V_{g0}}{v_{g1}}\right)^{2/3} \tag{13}$$

which means that: *the amplification obtained through regeneration equals the two-third power of the ratio of the "grid space" V_{g0} into the amplitude obtained with zero regeneration.* The much greater gain obtained through regeneration with small signals than with stronger ones is at once apparent by (13).

Some measurements provided a very satisfactory experimental verification of the theory outlined above.

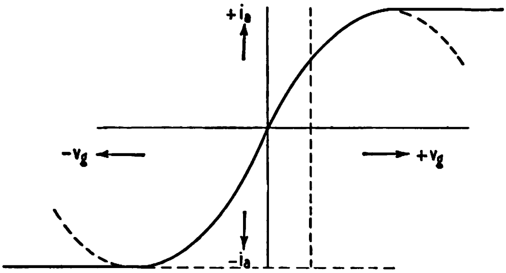

Fig. 2

The measurements were taken with relatively low frequency, thus avoiding obvious errors. That the results are, however, equally applicable to high-frequency circuits follows at once from the following theorem:

If a model is made of a high-frequency system consisting of linear inductances, linear capacities and non-linear resistances (e.g. triodes) and if the values of all the inductances (self and mutual) and capacities in the model are made n times these values in the original high-frequency system, but if the resistances (linear and non-linear) in the model are made equal to the resistances in the original circuit, the currents and potentials occurring in this model will be exactly equal in magnitude to the currents and potentials in the original high-frequency system but, considered as a function of the time, they will vary n times slower.[4]

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[4] Incidentally, in this model system the stray capacities are reduced n times in magnitude. Therefore, in order to investigate the effect of a specified stray capacity in the original high-frequency circuit it is only necessary to insert at its place in the model circuit a capacity n times the original stray capacity.

Therefore, the natural periods of the model will be n times those of the original system, and the building up and decay of currents in the model will also occur n times slower.

This simple theorem at once follows from a dimensional consideration of the coefficients such as r, ωL, $1/\omega C$, $M\omega$, occurring in the differential equations, because, e.g.

$$\omega L = \frac{\omega}{n} \cdot nL, \text{ etc.}$$

In fact, with a system having a natural frequency of 500 Hertz (cycles per second) and with critically adjusted retroaction it often took a minute for the free oscillation of the triode system to reach its final value, hence it was not easy to decide, before the external emf was applied, whether the retroaction was exactly critical or not. Therefore, an intermediate way was chosen and a model of a receiving set was made having a natural frequency of 15000 Hertz.

Four sets of readings were taken[5] with a standard triode tungsten filament $V_f = 4.0$ volts, saturation current 11 milliamperes, anode potential varying between 100 and 200 volts, negative grid bias varying between -4.5 and -7.5 volts.

The applied emf $E \sin \omega_1 t$ was varied between 2.10^{-6} and 10^{-2} volts, the resulting alternating grid voltage V_{g1} with no regeneration varied between 100.10^{-6} and 0.5 volts, and the resulting alternating grid voltage b with critical regeneration varied between ca. 0.2 and 4 volts. The exponent S in the formula

$$C = (V_{g0}{}^2 \cdot v_{g1})^s$$

which, according to the above theory, should be

$$s = 0.33$$

was found from the four sets of measurements to be

$$\begin{array}{rl} s = & 0.36 \\ & 0.36 \\ & 0.36 \\ & 0.32 \\ \cline{2-2} \text{mean: } s = & 0.35 \end{array}$$

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[5] The experiments and calibration of the necessary amplifiers were performed by Messrs. K. Posthumus and R. Veldhuyzen. Care was taken that the receiver did not react on the transmitter.

while the "grid space" V_{g0} calculated from these measurements was

$$
\begin{aligned}
V_{g0} &= 26 \text{ volts} \\
&\ 25 \\
&\ 20 \\
&\ 15 \\
\hline
\text{mean: } V_{g0} &= 22 \text{ volts,}
\end{aligned}
$$

as compared with the value

$$V_{g0}=27 \text{ volts}$$

obtained directly from the characteristic.

The experimental value for the exponent S fits in well with the theoretical value; provided only those parts of the characteristic are considered for which no oscillation hysteresis occurs. The representation of the i_a, V_g characteristic, by a cubical parabola, with three constants S_1, S_2, and S_3 only, obviously cannot yield very accurate values for V_{g0} obtained with a voltage swing of not more than 0.35 volts.

Therefore, there is no doubt that for all practical purposes the theory given above fits in well with experiment.

Finally, we give some practical figures calculated from formula (12) with the following data:

$$\frac{\omega L}{r}=40,$$

$$V_{g0}=27 \text{ volts.}$$

Electromotive force E working in grid circuit	Resulting alternating grid voltage with no regeneration (V_{g1})	Resulting alternating grid voltage b, with critical regeneration	Amplification obtainable through critical regeneration (C/V_{g1})
10^{-6} Volts	0.04×10^{-3} Volts	0.31 Volts	7700
10^{-5} "	0.4×10^{-3} "	0.66 "	1600
10^{-4} "	4.0×10^{-3} "	1.4 "	360
10^{-3} "	0.04 "	3.1 "	77
10^{-2} "	0.4 "	6.6 "	16

CHARACTERISTICS OF OUTPUT TRANSFORMERS*

By

J. M. Thomson

(Radio Engineer, Ferranti Electric, Limited, Toronto, Canada)

Summary—*The paper deals with the operating characteristics of the Output Transformer, which are developed in terms of the known speaker, tube and transformer constants. In the first part of the paper the general formula for the speaker current is developed and the effect of varying the transformer constants shown. The turn ratio of the transformer for maximum speaker current is considered in relation to the commonly used impedance ratio formula. The limitation of the impedance ratio formula is then pointed out and limits set for its general use. The general form of the current frequency characteristic for exponential horns and dynamic cone speakers is then obtained and a general method for matching the speaker to the output tube is given.*

In the latter part of the paper, curves are given to show the results obtained in the mathematical part of the paper. The curves also include the results of tests made to check the fundamental formula. The effect of the turn ratio on the form of the current frequency characteristic is shown and a method of using the turn ratio of the output transformer to match the speaker to the output tube is given. A perfect transformer is also compared with a good commercial transformer and the general effect of the leakage inductance and the self capacity of the transformer shown.

THE general use of output transformers to protect the speakers and at the same time balance the power input at the high and low frequencies has created a great deal of interest in the operating characteristics of these transformers. In this paper the operating characteristics are developed in terms of the known transformer, speaker and tube constants.

The following assumptions will be made with respect to the characteristics of the transformer, speaker and output tube.

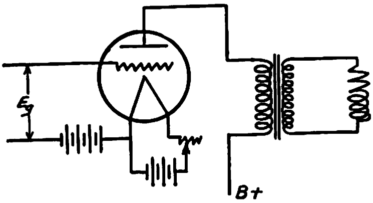

Fig. 1

The primary and secondary inductance are independent of frequency. This is true as long as the flux is uniformly dis-

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* Original Manuscript Received by the Institute, April 27, 1928. Presented before joint meeting of Buffalo-Niagara, Rochester and Canadian Sections at Buffalo, April 11, 1928.

tributed across the core. The inductance will vary with the amount of d.c. in the primary coils. It will also vary with the value of the a.c. voltage used in measuring it. However, for a fixed value of direct current and a given a.c. flux density the inductance will be essentially constant. Since the component of the self inductance due to the flux which passes through the coil itself is small, the unequal distribution of the current across the conductor will have very little effect on the inductance.

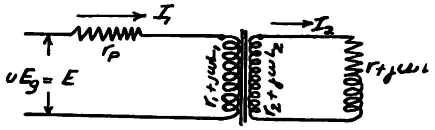

Fig. 2

The resistance of the primary and secondary coil is independent of frequency. The correctness of this assumption will depend on the size and shape of the conductors used and on the lengths of the leakage paths. This increase in resistance will

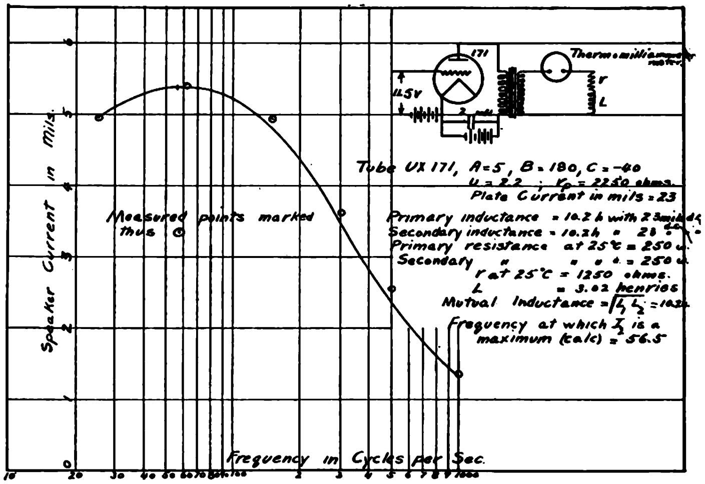

Fig. 3

usually be less than 5 per cent at 10,000 cycles, except possibly in very low ratio transformers made of few turns of relatively large wire.

The self capacities of the primary and the secondary coils are neglected.

The core loss has been neglected. It will be very small but could be represented by making the mutual inductance M a vector quantity instead of a pure number.

The amplification factor and the plate impedance of the tube will be assumed to be constant. The tube will also be assumed to be powerful enough to supply the required currents.

The two really unknown quantities are the speaker resistance and inductance. These are assumed to be independent of frequency. Actually they will vary a great deal with frequency. This in actual practice will necessitate measuring these quantities at each frequency and using the measured values in the formulas.

Let:

r_p = plate impedance of the tube in ohms
r_1 = resistance of the primary coil in ohms
r_2 = resistance of the secondary coil in ohms
r = resistance of the speaker coil in ohms
L = inductance of the speaker in henries
L_1 = inductance of the primary coil in henries
L_2 = inductance of the secondary in henries
M = mutual inductance in henries
U = amplification factor of the tube
$W = 2 \times 3.14 \times$ frequency in cycles per second
f = frequency in cycles per sec.
n = turn ratio of the transformer
I_1 = primary current
I_2 = secondary current
A = filament voltage in volts
B = plate voltage
C = grid bias in volts
E_g = a.c. voltage on the grid of the tube
$E = UE_g$
$j = \sqrt{-1}$
$L_e = L_2 + L$
$R_1 = r_1 + r_p$
$R_2 = r_2 + r$
$Z_1 = R_1 + jwL_1$
$Z_2 = R_2 + jwL_e$

The connection of the output transformer in the set is given in Fig. 1. Fig. 2 gives the equivalent circuit for alternating currents and voltages. The following currents were obtained by the use of Kirchoff's Laws.

$$E+jwMI_2=Z_1I_1 \tag{1}$$

$$jwMI_1=Z_2I_2 \tag{2}$$

On solving for I_1 and I_2 we get:

$$I_1=\frac{EZ_2}{Z_1Z_2+W^2M^2} \tag{3}$$

$$I_2=\frac{jwME}{Z_1Z_2+W^2M^2} \tag{4}$$

$$I_2=\frac{jwME}{(R_1+jwL_1)(R_2+jwL_e)+W^2M^2} \tag{5}$$

$$I_2=\frac{jwME}{R_1R_2-W^2L_1L_e+W^2M^2+j(R_1wL_e+R_2wL_1)} \tag{6}$$

$$I_2^2=\frac{W^2M^2E^2}{(R_1R_2-W^2L_1L_e+W^2M^2)^2+(R_1wL_e+R_2wL_1)^2} \tag{7}$$

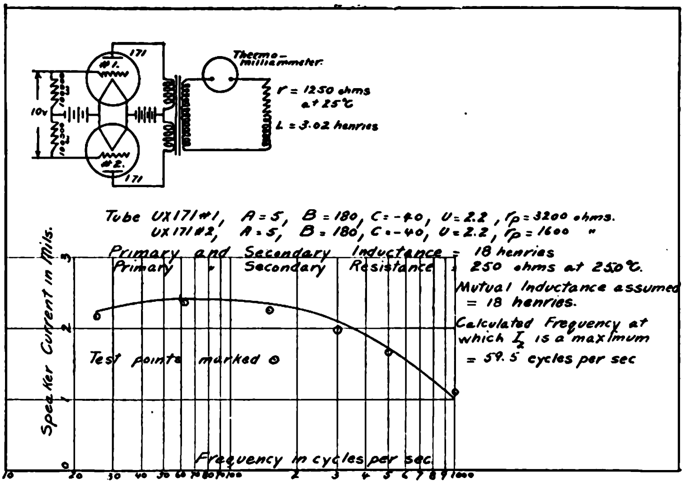

Fig. 4

The power input to the speaker will be a maximum when the current I_2 is a maximum, but I_2 is a maximum when I_2^2 is a maximum. Assume w to vary while R_1, R_2, L_1, L_2, L_e, M and E remain constant, then I_2 will be a maximum when

$$\frac{d}{dw}\left\{\frac{W^2}{(R_1R_2-W^2L_1L_e+W^2M^2)^2+(R_1wL_e+R_2wL_1)^2}\right\}=0 \tag{8}$$

From this we obtain:

$$W=\sqrt{\frac{R_1R_2}{L_1L_e-M^2}} \tag{9}$$

Let $L_2=n^2L_1$ and $M=nL_1$ (10)

then

$$W=\sqrt{\frac{R_1R_2}{L_1L}}$$

This is the frequency at which the current in the speaker will be a maximum. It is independent of the turn ratio "n" but depends directly on the value of the primary inductance. When L_1 is infinite the transformer is perfect.

$$W=0 \tag{11}$$

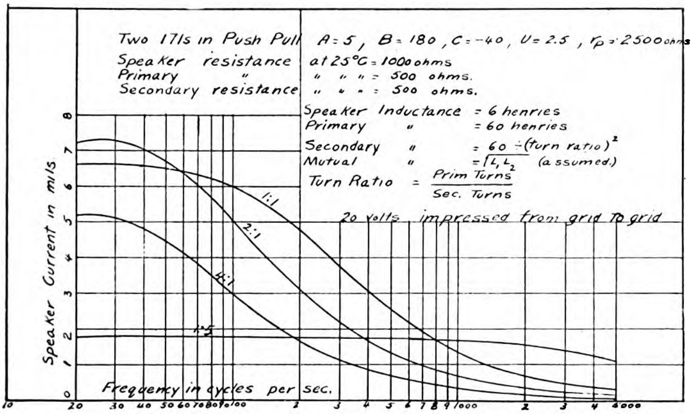

Fig. 5—Effect of Changing the Turn Ratio of the Output Transformer.

The current I_2 is limited only by the equivalent resistance of the system. Actually, at zero frequency the transformer will not function and I_2 will be zero. However, as long as the frequency is not zero the transformer will operate and a current will flow. This current will increase as the frequency is decreased until the limiting value as determined by the total equivalent resistance is reached. From then on the current will be constant.

When $L=0$

$$W=\alpha \tag{12}$$

and the current is a maximum for all values of W that make WL_1 very large as compared with R_1.

Assume $M=\sqrt{L_1L_2}=nL_1$ and $L_2=n^2L_1$ in (7), which indicate that the coefficient of coupling between L_1 and L_2 is unity. The leakage inductance in a good transformer should be less than ½ of 1 per cent. On making the above substitution in (7), we get:

$$I_2^2=\frac{w^2n^2L^2E^2}{(R_1R_2-w^2L_1L)^2+(R_1wL+R_1wL_1n^2+R_2wL_1)^2} \qquad (13)$$

Assume n to vary while R_1, R_2, L_1, w_1, L_1, M, and E remain constant. Then I_2 is a maximum when

$$\frac{d}{dn}\left\{\frac{n^2}{(R_1R_2-w^2L_1L_2)^2+(R_1wL+R_1wL_1n^2+R_2wL_1)^2}\right\}=0 \qquad (14)$$

On solving we get:

$$n^2=\sqrt{1+\frac{R_1^2}{w^2L_1^2}}\times\sqrt{\frac{R_2^2+w^2L^2}{R_1^2}} \qquad (15)$$

Equation (15) will give us the value of n for which the current I_2 is a maximum. Since in general n varies with w, each frequency will require a different turn ratio to give the maximum input. When $L=0$ and $\left(\frac{R_1}{WL_1}\right)^2$ is very much smaller than 1, n^2 will be equal to R_2/R_1 and is then independent of frequency. In a properly designed transformer n^2 will be equal to the ratio of the secondary and primary impedances of the transformer. Therefore the impedance ratio of the transformer windings is equal to

$$\sqrt{1+\frac{R_1^2}{w^2L_1^2}}\times\sqrt{\frac{R_2^2+w^2L^2}{R_1^2}}$$

This ratio will depend on the value of the primary reactance as compared with the total resistance in the primary circuit. When L_1 is very large, $1+\frac{R_1^2}{w^2L_1^2}$ becomes equal to one and

$$n^2=\sqrt{\frac{R_2^2+w^2L^2}{R_1^2}} \qquad (16)$$

This is the commonly used impedance ratio rule and it is generally stated as follows:

The current I_2 will be a maximum when the impedance ratio of the transformer windings is equal to the impedance of the secondary load divided by the plate impedance of the tube. This is incorrect unless the primary resistance of the transformer is added to the plate resistance and the secondary resistance is added to the load resistance. In any case this rule fails as soon as wL_1 begins to approach r_p in value. When this occurs (15) must be used to obtain the correct turn or impedance ratio.

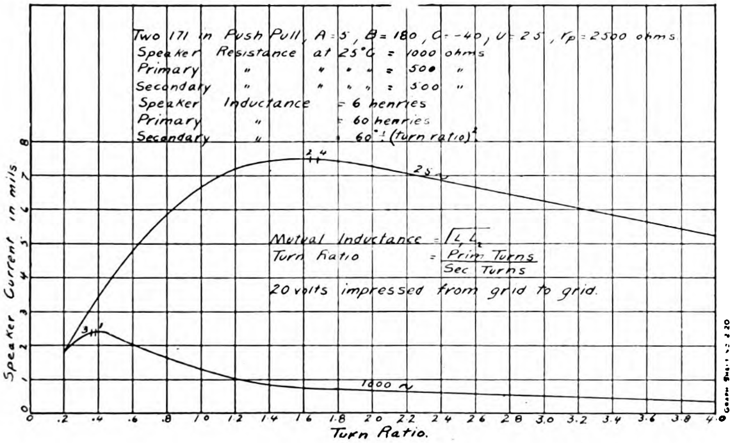

Fig. 6

In (13) assume L_1 to vary while w, n, E, R_1, R_2 and L remain constant. I_2 will be a maximum when

$$\frac{d}{dL_1}\left(\frac{L_1^2}{(R_1R_2-w^2L_1L)^2+(R_1wL+R_1wL_1n^2+R_2wL_1)^2}\right)=0 \qquad (17)$$

Solving we get

$$L_1=-\frac{R_2^2}{w^2n^2L}-\frac{L}{n^2} \qquad (18)$$

This means that the transformer primary inductance must have the properties of a condenser.

When $L=0$

$$L_1=\alpha \qquad (19)$$

When W^2L is very large as compared with R_2^2

$$L_1 = -\frac{L}{n^2} \tag{20}$$

To obtain this condition in practice would require a transformer with excessive self capacity in the primary and secondary coils, and this would lead to a cut-off of the higher frequencies.

To obtain an idea as to what values of capacity are required, assume $L = 6$ henries and $n = 1$, with $f = 50$ cycles per second. Then $wL_1 = -6 \times 314 = -1880$ ohms. The capacity required to give this will be equal to 1.69 μf. This would act as a short circuit for all frequencies above 500 cycles.

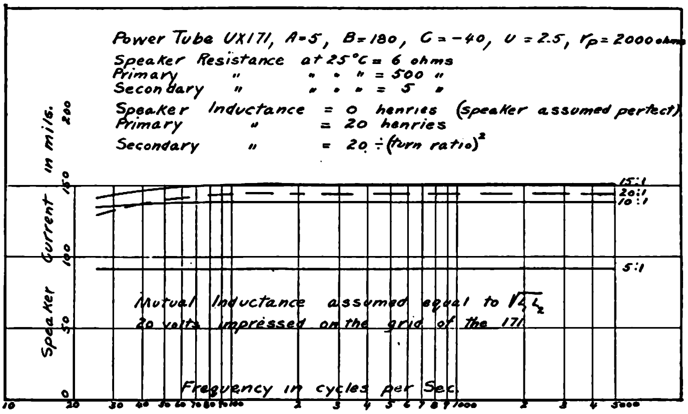

Fig. 7—Effect of Changing the Turn Ratio of the Output Transformer with Inertia-Controlled, Electro-Dynamic Cone Speakers.

The following curves have been plotted to show visually some of the results obtained in the mathematical part of the paper.

The curves in Figs. 3 and 4 give the results of tests made to check the basic formula for I_2 as given in equation (6). The test points are substantially in agreement with the calculated. The frequency for maximum I_2 as calculated by (10) is marked on the curves. As the curve is quite flat at these points it is rather difficult to pick out the frequency at which the current is a maximum. The leakage inductance of the transformer used in the tests of Fig. 3 was less than 0.0025 henries. In spite of the bad unbalance in the tubes used in the push-pull tests of Fig. 4, the calculated currents are a close approximation to the test results. The maximum error was less than 10 per cent. This test was made to determine the limits that must be set for the leakage inductance in a good transformer.

The remaining curves give calculated results only but the approximate tube conditions are marked on each curve.

Figs. 5 and 6 give the effect of varying the turn ratio for speakers of the ordinary exponential horn and cone types. It is easily seen from these curves why the power tube overloads so easily when amplifiers which bring in the low notes are used.

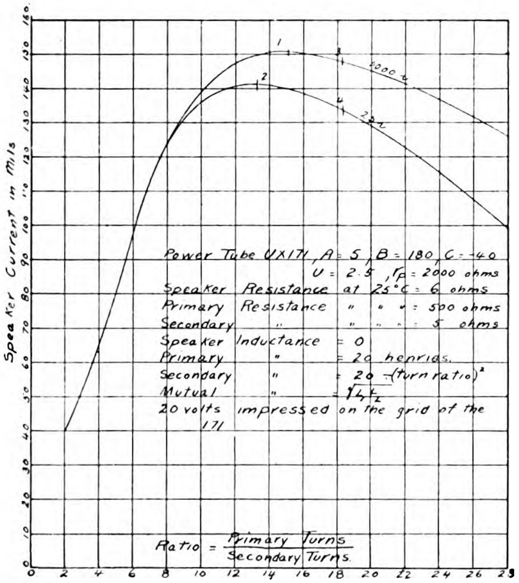

Fig. 8

For example, in the 1:1 ratio curve of Fig. 5 the current at 30 cycles is 4.9 times the current input at 1000 cycles. With the 2:1 ratio the current ratio for the same frequencies is 11:1. The ratio of the power at the lower frequencies can thus be changed within wide limits. But since the frequencies that give the sense of volume lie between 200 to 2500 cycles, it will be found that as the lower frequencies are increased the higher frequencies are decreased and the volume will fall off. These curves can be used to match the speaker to the tube. The speaker characteristic required for this is the current input required to give constant response from the speaker at the different frequencies. The average curve neglecting the pronounced resonant peaks must be used. The transformer ratio which gives a current frequency characteristic similar in form to the speaker characteristic is the correct one to use. It is not usually possible to make the transformer correct the resonant peaks of the speaker. Special filter systems are usually required for this.

The curves in Fig. 6 give the speaker currents at 25 and 1000 cycles for different ratios. Points 1 and 2 give the turn ratio as calculated by (15), while points 3 and 4 give the turn ratio as calculated by the usual impedance ratio method. As L_1 is large, the two results are close, with (15) giving the correct result.

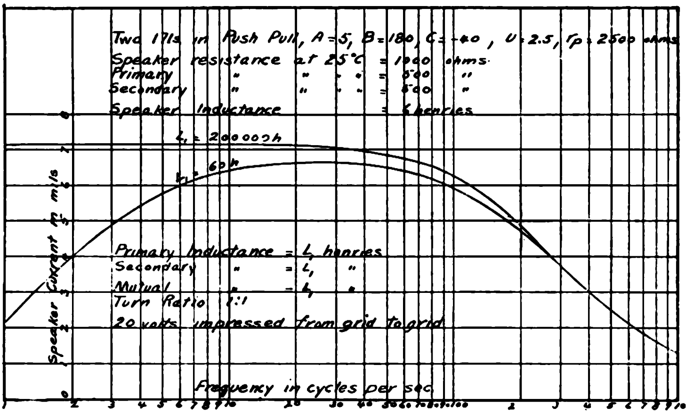

Fig. 9—Effect of Changing the Primary Inductance of the Output Transformer.

Figs. 7 and 8 give the results obtained with perfect speakers of the electro-dynamic cone type. These curves also show the need for power tubes when speakers of this type are used. The current frequency characteristic required for the perfect speaker of this type is a straight line with the current input independent of frequency.

Fig. 8 gives the turn ratio for maximum speaker current at 25 and 1000 cycles. The correct ratio to use with a perfect speaker will be 8. This will give a speaker current of 124 mils independent of frequency. Points 1 and 2 give the turn ratio as calculated by equation (15). Points 3 and 4 give the turn ratio as calculated by the impedance ratio method. The error in the impedance ratio method is greater in this case because wL_1 is not very much greater than r_p and the secondary resistance of the transformer is almost equal to the speaker resistance. This would be a very inefficient transformer. With this type of speaker the secondary resistance should be less than 1/10 of the speaker resistance.

The curves of Fig. 9 show the difference between a good transformer and a transformer with almost perfect primary inductance. The commercial product at 10 cycles is 90 per cent as efficient as the perfect transformer. At 30 cycles it is 94 per cent efficient.

The curves of Fig. 10 show the effect of the primary inductance on the turn ratio. For very low values of inductance

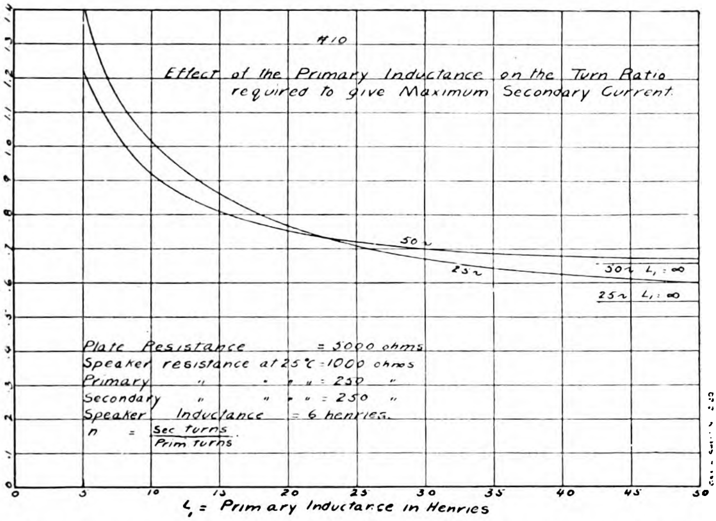

Fig. 10—Effect of the Primary Inductance on the Turn Ratio Required To Give Maximum Secondary Current.

the impedance ratio method may be as much as 100 per cent in error.

The curves of Fig. 11 show the effect of the leakage inductance of the transformer on the shape of the current frequency characteristic. The curve for M59 would correspond to a very poor commercial transformer. The curve marked M59, $C\text{-}100 \times 10^{-12}$ farads shows the combined effect of the leakage inductance and the self capacity of the secondary coil. In general at the higher frequencies the self capacity tends to neutralize the effect of the leakage inductance.

In conclusion some idea should, perhaps, be given of the use of the developed formulas. Knowing the speaker characteristic, the primary inductance is first chosen so as to make at least the maximum current occur at the lowest frequency desired. The turn ratio is then varied until the form of the speaker and the transformer current frequency characteristics are the same.

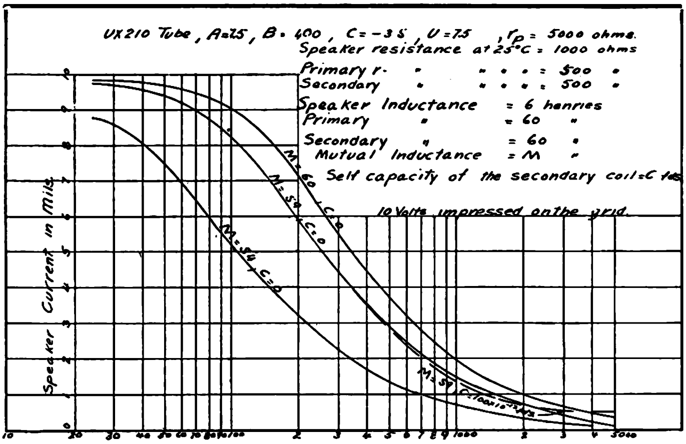

Fig. 11—Effect of Leakage Reactance on the Current in the Speaker.

It must be emphasized that the turn ratio has no effect on the frequency at which the current is a maximum. This will depend on the total resistances in the primary and secondary circuits and on the inductance of the primary and speaker.

CONCLUSION

Operating characteristics of output transformers used to protect the loudspeaker and to balance the power at various frequencies are developed in this paper.

The necessary conditions for maximum current in the speaker are $\omega=\sqrt{\dfrac{R_1R_2}{L_1L}}$ with the frequency varying.

With the frequency constant, I_2 is a maximum when,

$$n^2=\sqrt{1+\frac{R_1^2}{w^2L_1^2}}\times\sqrt{\frac{R_2^2+w_2L^2}{R_1^2}}$$

When wL_1 is very large this reduces to $\sqrt{\dfrac{R_2^2+w^2L^2}{R_1^2}}$ and this is known as the impedance ratio formula. Test and calculated curves are then given to show the effect of the primary inductance, turn ratio, leakage inductance and self capacity on the current-frequency characteristic of the transformer.

COOPERATION BETWEEN THE INSTITUTE OF RADIO ENGINEERS AND MANUFACTURERS' ASSOCIATIONS*

By

ALFRED N. GOLDSMITH†

(Chief Broadcast Engineer, Radio Corporation of America, New York City)

IT is at once a real privilege and an unusual opportunity to address the Radio Division of the National Electrical Manufacturers' Association on the subject of cooperation between the Institute of Radio Engineers and the organizations representing radio manufacturers. In the early stages of the development of any new industry, and its corresponding division of technology, there is frequently little understanding or cooperation between the engineer and the industrial executive, or between the associations of which these two are members. Everyone is too busy dreaming, planning, developing, manufacturing, selling and servicing to have much time to consider effective methods of cooperation with members of the other group. As a result, industry proceeds for a while along comparatively disorderly lines, imperfectly controlled by sound engineering principles. And, on the other hand, the engineer (perhaps somewhat discouraged by an apparent lack of interest in what he can contribute to the development of the industry) is less effective in producing new ideas and having them commercially applied than would be the case under more favorable auspices.

Fortunately, radio has now reached a stage of development where the workers in that interesting field occasionally have a chance to "take stock" and to lay out broad plans for the systematic and effective development of their art by closer cooperation between the engineers and the industrial officials and executives.

I appreciate very much the invitation to address you on behalf of the Institute of Radio Engineers. At the risk of needless repetition (since I have no doubt that most of you are well-acquainted with the nature of the Institute and its activities) I will only say that it is a scientific society of international character, having as its object the advancement of the science and art of radio communication. It numbers among its 4500 mem-

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† President of the Institute, 1928.

bers a large percentage of the leading engineers and workers in the world scattered from Singapore to Saskatchewan and from Paris to Portland. It encourages progress in radio engineering through the publication of engineering and scientific papers, through cooperation with other scientific societies, through the award of prizes for especially meritorious advances in radio, through standardization activities and in other ways. Its principal publication, the PROCEEDINGS of the Institute of Radio Engineers, has the distinction of having presented in its pages the original descriptions of many of the greatest radio advances of the last twenty years and of being a standard reference book to radio engineers and students alike. Its Board of Direction is continually cognizant of the importance of maintaining its reputation as a disinterested scientific body. In its activities it endeavors to function as an unbiased organization fostering the general interchange of sound engineering information.

In the United States, the Institute of Radio Engineers, with its headquarters in New York, has Sections in Atlanta, Boston, Buffalo, Chicago, Cleveland, the Connecticut Valley, Detroit, Los Angeles, New Orleans, Philadelphia, Pittsburgh, Rochester, San Francisco, Seattle, and Washington. There is a vigorously active Canadian Section with headquarters at Toronto.

It is the desire of the Board of Direction of the Institute that the facilities and capabilities of the Institute be fully utilized in friendly cooperation with manufacturing associations to the end that the radio art may be more rapidly advanced. My remarks today have been assembled with due regard to the viewpoints of the members of the Board of Direction of the Institute rather than as an original contribution. They have suggested many points in connection with the following four principal types of cooperation which the Institute of Radio Engineers can work out with the manufacturing associations.

I. MEETINGS OF THE INSTITUTE AS AN INFLUENCE IN THE TRAINING OF ENGINEERS

It is self-evident that prospective radio engineers gain a valuable type of training and education through attendance at Institute Meetings and through study of its publications. These young engineers are sooner or later employed by the member companies of manufacturing associations and any information they have gained through listening at meetings of the Institute, through articles in the PROCEEDINGS, and through contact with other engineers at the meetings is an intangible but nevertheless real asset of the companies which employ them.

It is therefore a reciprocal responsibility of the member companies to permit the timely release of information on new radio methods and devices originating in their organization through the presentation of papers before the Institute of Radio Engineers since in this way the technical information of today becomes the capital of the engineers of tomorrow who, employed by the various manufacturing companies, will return in full measure the benefits of their training. Manufacturing associations should therefore encourage to the utmost the presentation of papers by their member companies on new radio equipment and method of operation.

II. MUTUAL INTEREST IN THE PUBLICATION OF CONTRIBUTIONS TO ENGINEERING KNOWLEDGE

As has already been intimated, most of the important radio advances during the last twenty years have been described in the PROCEEDINGS. The practicing engineers who have read these pioneer articles have benefitted from them and they have helped to keep the organization with which they were connected in the vanguard of progress.

The Institute prints and distributes a great deal of information of interest to manufacturers themselves in the PROCEEDINGS, in its Standards Hand Book, its Year Book, and its Index. It is expected that these activities will show a healthy and steady expansion and that the manufacturers and their engineers will be able to gain increasing practical benefit from the material in the Institute publications. Among the more recent interesting offshoots along these lines is the publication of features of general news interest together with the notes on the activities of the various sections and of individual members of the Institute.

The radio manufacturers have the duty and privilege of encouraging, even at the cost of time and money, the preparation of technical papers by their research and engineering staffs. The Institute has the duty of publishing this material and of thus making it available to the world. The issuance of the PROCEEDINGS on a monthly basis rather than bimonthly as formerly is of definite value in accomplishing this.

This Institute of Radio Engineers, in endeavoring to publish reputable and sound technical information from whatever source, is following a policy which must certainly have the support of the manufacturers' associations.

Responsible radio manufacturers have consistently avoided disclosures of supposedly marvelous but actually unproven developments in the radio field. They have shown by the truthful and conservative character of their publications descriptive of new equipment or methods that they appreciate the injurious effect upon public confidence of premature and exaggerated descriptions of trivial or inconsequential modifications of existing methods or equipment. The leading radio manufacturers of the United States have frowned upon every instance of technical journal or newspaper publicity which corresponds not to actual accomplishment in the field of service to the public or scientific and engineering advancement but to occasional, imperfect or doubtful results. The Institute of Radio Engineers is heartily in accord with this policy and may be depended upon to proceed on the basis of a similar code for publications.

III. Mutual Interest in Standardization

There is available to all the experience of the Institute of Radio Engineers in the formation of the language of radio. For nearly twenty years the Institute has been working at this task of formulating definitions of terms and the standardization of symbols, and has published one edition after another of its standards during the last fifteen years.

Quoting part from a publication of the Institute, I may say that during the course of 1927 there has been some discussion of the relation between the standardization work of the Institute of Radio Engineers and that of the associations of radio manufacturers. The chairman of the Institute Standardization Committee has had some correspondence with a number of persons interested in this question and has found that the following appears to meet with general acceptance as an expression of the relationship between the standardization work of these two groups. It is recognized, of course, that in this new field it is impossible at the present time to determine upon any hard and fast dividing line.

Institute of Radio Engineers:

(1) Terms, definitions and symbols, and

(2) Methods of testing materials and apparatus in order to determine their important characteristics. This work may consist of purely advisory discussion as to convenient forms of tests, precautions to be taken, etc., or it may include standardization of definite test procedures to serve as a common basis of comparison of the properties of performance of material or apparatus.

Manufacturers' Groups:

(1) Standardization of size and characteristics of apparatus, to promote interchangeability of parts, either mechanical or electrical, and

(2) Setting of standard ratings for the properties or performance of material or apparatus.

While the above statement has never been formally adopted the general principles which it expresses are being followed by the several organizations concerned. Although the dividing line between the several fields of work cannot be rigidly drawn, it is believed nevertheless that the above classification will help in guiding the efforts of the committees which are working on various aspects of radio standardization.

As has been indicated above, the Institute is active in the formulation of methods of testing materials and equipment. The importance of such an activity to the manufacturer is obvious. Success and progress depend upon a definite knowledge of the performance of the raw materials which go into apparatus or equipment as well as the performance of the finished article. In developing technically correct and readily usable standard test methods, the Institute of Radio Engineers is helping the manufacturers' associations. The manufacturers, on the other hand, can reciprocate by adopting as standard the test methods developed by the Institute in relation to radio apparatus and circuits. In doing this, the manufacturers can adopt any limits or tolerances required for a particular product or particular performance using the Institute test methods.

It should be added that the Institute of Radio Engineers and the American Institute of Electrical Engineers are the sponsor bodies of the Sectional Committee on Radio organized according to the procedure of the American Engineering Standards Committee and that the Sectional Committee is entrusted with the task of recommending the final American standards to the American Engineering Standards Committee. There are therefore no organization difficulties in the way of as large a measure of radio standardization as is consistent with the growth and development of the art.

IV. MUTUAL INTEREST IN TECHNICAL PROBLEMS OF RADIO REGULATION

The Institute stands ready, through the appointment of committees, or by otherwise utilizing its facilities, to cooperate in the study of various technical problems of general interest to its membership. For example, the Institute of Radio Engineers sent representatives to the several National Radio Conferences held at the call of the Department of Commerce in Washington. These representatives participated in the formulation of the recommendations which, informally adopted as guides by the Department of Commerce in the administration of the law, were markedly instrumental in the rapid development of the radio art. It is believed that one of the most important fields for cooperation between the Institute and other parts of the radio industry is in the technical aspects of radio regulation. The study and formulation of the requirements which must be met in the allocation of frequencies to stations and services are of fundamental importance if interference is to be kept at a minimum in the crowded radio spectrum.

The problem of securing a maximum of radio service with minimum of interference but without excessive cost is a matter of mutual interest to engineers, manufacturers, operating agencies, and the public. While the public does not, of course, care so much about the mechanism as it cares for satisfactory reception and radio service generally, the other agencies are interested in the various elements which contribute to the correct solution of this sort of a problem.

It is clear that it is generally advantageous for the Federal regulatory body entrusted with the enforcing of the law to consult with the manufacturers and the radio engineers before reaching definite conclusions. It is therefore important that the manufacturers and the engineer shall cooperate to the end that sound engineering principles may be supplemented by full knowledge of the conditions of the industry and that feasible recommendations shall emanate from both the manufacturing and engineers' groups.

An instance of this sort occurred recently. The Federal Radio

Commission, as you know, in its effort to disentangle the threads of radio broadcasting allocation, invited the Institute of Radio Engineers, the broadcasters, the manufacturers' associations and others to present their suggestions as to a concrete plan for the betterment of broadcasting service to the public, to be consistent with the existing law. It is greatly to be hoped that with the constructive cooperation of the entire industry, the Commission will be able to work out a solution, which is at the same time practical and based on sound technical considerations.

The Institute has furnished information to the Federal Radio Commission on other matters, from time to time, and it desires the closest possible contact with the manufacturers' association to the end that the information which it furnishes shall be not only scientifically correct but economically feasible and capable of ready application. If the manufacturers' associations will appoint representative committees to consider matters involving Federal radio regulation with corresponding committees of the Institute there can be no doubt that there will be much wisdom in the combined recommendations.

Some additional concrete recommendations which I should like to make are that members of manufacturers' associations should become members of the Institute of Radio Engineers, that they should place the publications of the Institute before their employees, and that they should look to the publications of the Institute of Radio Engineers for some assistance in the solution of the problems which confront them, and that they should encourage their employees to become active members of the Institute committees. The manufacturers' associations will do well to remember that what they sell to the public is nothing more than a mixture of raw materials and engineers' brains.

One final important aspect of the relationship between manufacturers' associations representing the radio industry and the Institute of Radio Engineers is its ever-continuing nature. Problems once tentatively solved immediately arise again with new aspects and with new interests injected into them. Cooperation is not a thing of today and tomorrow, it is an unending task, a fundamental need, and a provider of rich rewards both in public esteem and commercial success.

NOTES ON QUARTZ PLATES, AIR GAP EFFECT, AND AUDIO-FREQUENCY GENERATION*

By

August Hund

(Bureau of Standards, Washington, D. C.)

Summary—*Experiments on the frequency of piezo-electric elements are described with special reference to the effect due to supersonic sound waves generated in the air gap of the holder and due to its capacity. It is shown that a mechanical load on the crystal increases its thickness frequency and that an air gap has a similar effect. The velocity of the supersonic sound waves is about the same as for ordinary sound waves. The value found is 338.68 meters per second at 24.5C deg. An appropriate air gap gives even more high-frequency output than a mechanically-loaded crystal and procures a steady frequency operation. Two sputtered piezo-electric elements can produce a beat frequency which is correct within a few parts in 100,000. A method is shown by means of which a low-frequency standard can be obtained by harmonic division of a high frequency due to piezo-electric element.*

ANYONE experimenting with piezo-electric plates will notice that the temperature, the load in the anode branch, and the holder of the piezo-electric element affect somewhat the frequency of the output current.

The effect is very pronounced when the beat frequency of two vibrating crystals is to be taken as a standard since the percentage error in the beat frequency is larger in the ratio

$$\frac{\text{average value of the two high frequencies}}{\text{beat frequency}}$$

than the percentage error in any one of the two high frequencies.

The temperature effect can be checked by a suitable thermostatic control and the circuit effect by using a fixed anode load.

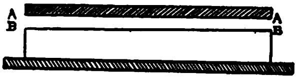

Fig. 1—Piezo-Electric Element and Its Electrodes.

The holder effect which is the subject of this discussion is twofold: One being due to the supersonic sound waves displaying

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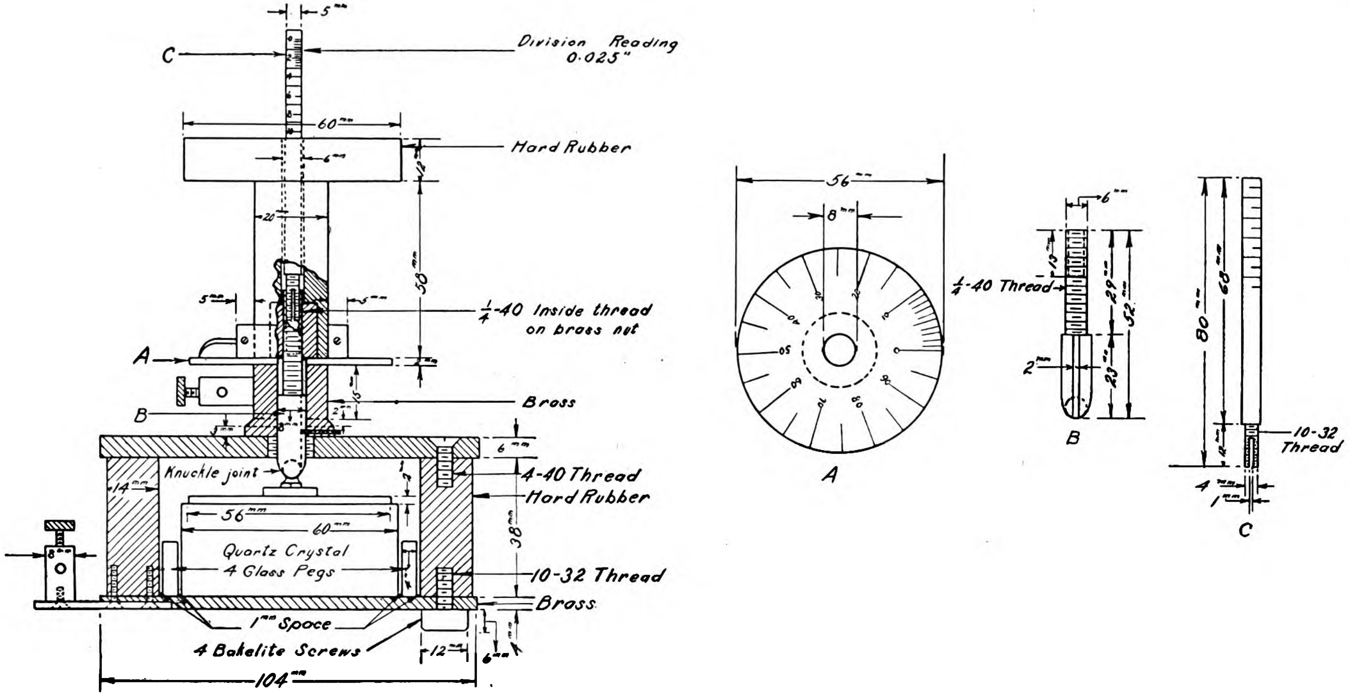

Fig. 2—Piezo-Electric Crystal Holder with Micrometer Adjustment.

themselves between the two reflecting walls A-A and B-B (Fig. 1) and the other due to the capacity of the crystal holder. The first one can be avoided by putting the holder with the piezo-electric element under a vacuum and the other can be kept fixed by using metal coated crystals. When a contact brush is used as the upper electrode and the crystal rests on a metal plate, operation can be obtained which gives beat frequencies accurate to a few parts in a hundred thousand. It will be found advantageous to work the two high-frequency currents into a shielded grid tube so that the back action of the load is practically avoided.

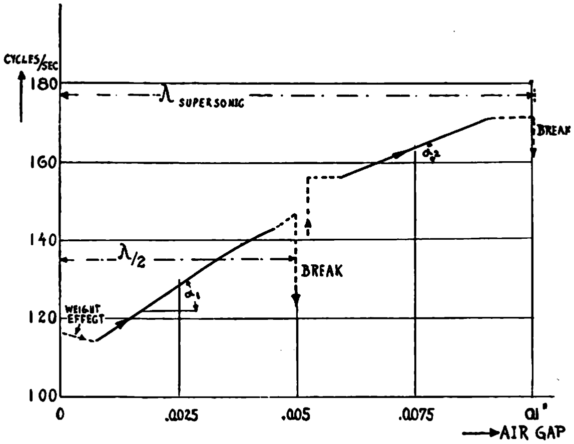

Fig. 3—Frequency Change with Air Gap.

In the course of this study the following was observed: The true crystal frequency is never used. It is either

(a) "the metallized crystal" frequency (crystal-coated with two metal layers), or

(b) the "crystal-crystal holder" frequency, or

(c) the latter partly affected by the load at a certain temperature.

Putting a mechanical load such as the metal electrode on the piezo-electric element increases its frequency, removing the electrode very gradually produces for a moment the true crystal frequency, while a very small air gap gives again a somewhat higher frequency. Fig. 2 indicates one of the plate holders with which the above effects were studied.

Fig. 3 verifies some of the above statements. For this case two large disk crystals were excited in the thickness vibration (frequency about 130 kc per second) and the output branches coupled to a common coil which had a rectifier in the circuit. The ordinates give the beat frequency and the abscissas the air-gap distance in inches.

It is seen that with the upper electrode just resting on the piezo-electric element the frequency is about 118 cycles per second. It drops then to a value of about 116 cycles per second and ascends again as indicated. The oscillation suddenly stops since the condition for one-half wavelength of the supersonic sound wave exists, and from the λ/2 as well as from the λ condition, the sound velocity in the gap is found as $v = 0.258064 \times 131240 = 338.68$ meters per second at 24.5 deg. C. This shows that the velocity of supersonic sound waves is about the same as found for audible sounds as was also found by G. W. Pierce.[1]

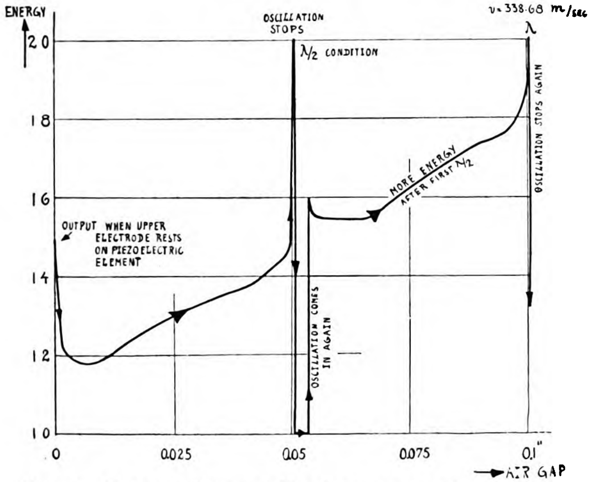

Fig. 4—Energy Curve with Respect to the Air Gap.

After the λ/2 condition the oscillation starts in again and proceeds along a slope α_2 somewhat different than before. Many other crystals show about the same behavior.

Fig. 3 indicates that it is not wise to work with such a small air gap that the upper electrode nearly touches the piezo-electric element or with a gap distance very close to λ/2, λ, 3/2λ, etc., of the supersonic sound wave. This is also shown by the curves of Fig. 4 where the ordinates stand for the energy output and the abscissas for the air gap. A very good position seems to be that which provides an air gap of 3/4 of the supersonic sound waves. Fig. 4 is a typical example of many tests of this kind.

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[1] *Proc.* American Acad. of Science, **10**, 271; 1925.

The settings with an air gap are no doubt of value for high-frequency work while the metal sputtered piezo-electric elements seem at the present time most promising for low-frequency work.

When, therefore, a metal-coated piezo-electric element is used the accuracy of the high frequency must be very great, and a subharmonic in the audio-frequency range should give an audio-frequency standard which is just as accurate as the high frequency. One of the best methods of producing subharmonics is indicated in Van der Pol's paper on Relaxation Oscillations.[2] Applying such oscillations to our case suggests the new method for piezo-electric frequency standardization as indicated in Fig. 5. The voltage $E\epsilon^{j2\pi ft}$ is taken from a piezo-electric oscillator utilizing sputtered metal electrodes. The condenser C is varied and the output will produce audio-frequency currents which are sub-multiples of the high-frequency voltage $E\epsilon^{j\omega t}$ applied to the gas discharge system. Only frequencies which are submultiples will be given off at the output end. As the condenser C is varied the tones observed in the output branch will vary in steps producing the frequencies $f, f/2, f/3, f/4$, etc. The system will work just as well when the terminals for $E\epsilon^{j\omega t}$ are short-circuited and the emf due to the piezo-electric element is applied across the condenser C. A multivibrator and other arrangements using relaxation oscillations whose period is roughly given by $T = \pi/2 \cdot CR$ can also be used for frequency division.

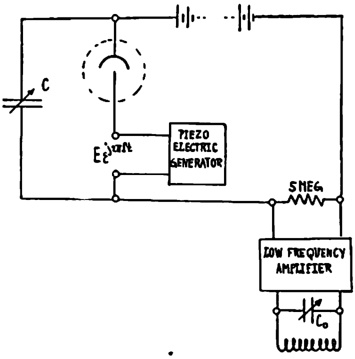

Fig. 5—Stepping Down of the Frequency of a Piezo-Electric Element.

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[2] Z. f. Hochfrequenztechnik, Dec. 1926, 178–187; April, 1927, 114–118.

NOTES ON APERIODIC AMPLIFICATION AND APPLICATION TO THE STUDY OF ATMOSPHERICS*

By

August Hund

(Bureau of Standards, Washington, D. C.)

Summary—*It is shown that ordinary amplifiers cannot produce true repeater action when aperiodic transients are impressed on the system. The output current can, however, be interpreted in terms of the coupling. For true aperiodic amplification the time constants have to be chosen such that the system gives true repeater action. An aperiodic amplifier using space-charge tubes is shown.*

That a certain type of an amplifier produces harmonic output currents for harmonic voltages of the same frequency is by no means a proof that the system would give correct repeater action in all cases, as, for instance, when aperiodic discharges, study of atmospherics and the like are to be investigated with respect to their shape.

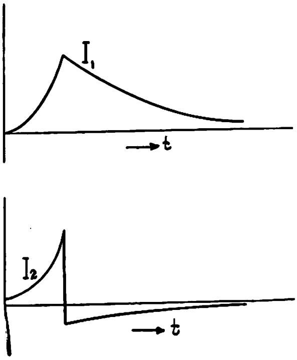

Fig. 1—Current-Time Curves.

A transformer-coupled amplifier produces, for instance, a voltage variation in each succeeding stage which depends on the time derivative of the current flowing in the primary of the transformer while for a condenser coupling the time integral of the current causes a change of the shape of an aperiodic transient. When sinusoidal variations are amplified the wave

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shape remains the same since only a phase shift (of 90 degrees for no losses) takes place and the output current such as noted with a suitable oscillograph will give a picture of the true display.

Fig. 1 gives a characteristic aperiodic curve I_1, as atmospherics have sometimes been shown in the literature while the I_2 curve is the time derivative of it to some arbitrary scale. The upper curve can therefore also be looked upon as being due to the time integral of the I_2 curve.

It is therefore evident that the shape of the output current has to be interpreted when such effects occur either along an amplifier or are due to a coupling to an oscillograph.

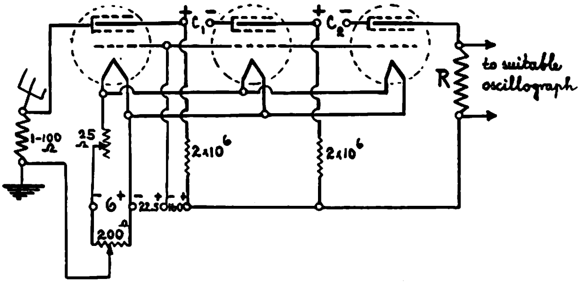

Fig. 2—Aperiodic Amplifier.

There are several means for preserving the original shape of the transient. One method is to use such time constants that almost true repeater amplification prevails, and no noticeable change in the shape occurs. Another means is to use a straight resistance-coupled amplifier such as is shown in Fig. 2, where UX-222 tubes or their equivalent are used in the space charge connection. In many cases two stages of amplification are sufficient. When an ordinary oscillograph is used a UX-210 or its equivalent may be employed in the last stage.

EFFECT OF THE ANTENNA IN TUNING RADIO RECEIVERS AND METHODS OF COMPENSATING FOR IT*

By

Sylvan Harris

(Engineering Laboratory, Brandes Products Corp., Newark, N. J.)

Summary—*The reflected effect of the primary circuit of a resonance transformer, or of the antenna circuit of a radio receiver, upon the secondary coupled to it, is discussed. Equations are given for determining the apparent change of inductance in the secondary and methods of compensating for these changes are given.*

It is a common practice in designing radio receivers to make the inductance of the antenna or input circuit so small that the resonance frequency of that circuit is higher than any frequency to which the receiver is to be tuned. Moreover, the mutual inductance of the input coupling transformer is kept small, so that the effect of the primary (antenna) circuit upon the tuning conditions of the secondary circuit coupled to it is as small as possible. By so doing, it becomes possible to build radio receivers in which the circuits of the cascade amplifier can be tuned in unison by a master control.

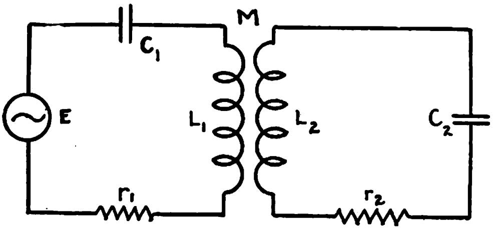

Fig. 1

It is clear, however that much of the ability of the amplifier to absorb power from the antenna circuit is lost; besides, there is always sufficient coupling between the circuits to cause detuning of the first stage. The detuning effect is more marked the higher the frequency, and would be very considerable at or near the resonance frequency of the primary or antenna circuit.

It is the purpose of this paper to study the effects of the antenna circuit upon the secondary circuit, to derive equations whereby these effects may be calculated, and to present several

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* Original Manuscript Received by the Institute, April 24, 1928.

methods of compensating for these effects. The theory employed applies rigidly to the resonance transformer circuit of Fig. 1, and also applies to an antenna coupling system insofar as the antenna may be regarded as equivalent to a lumped inductance in series with a lumped capacity. The discrepancy, however, is small enough to neglect for all ordinary purposes, especially since any errors which might arise are mitigated by the variable nature of the compensators to be described. For simplicity, therefore, the discussion will be centered around the circuits of Fig. 1.

Apparent Change of Inductance

Since, in the circuits usually employed all the tuning is done by adjustments of the secondary constants, maximum current is obtained in the secondary circuit when the following relation exists:

$$\frac{x_2}{x_1}=\frac{\omega^2 M^2}{z_1^2} \tag{1}$$

This relation is termed by Pierce[1] "Partial Resonance Relation *S*," and may be written in the form

$$\begin{aligned}\frac{1}{\omega C_2}&=\omega\left[L_2-\frac{\omega M^2}{z_1^2}x_1\right]\\&=\omega[L_2-\Delta L_2]\\&=\omega L_a\end{aligned} \tag{2}$$

Eq. (2) indicates that the apparent secondary inductance L_a differs from the true inductance L_2 by the amount ΔL_2, and whether it is greater or smaller than the true value depends upon the sign of x_1. The complete expression is

$$\Delta L_2=\frac{\omega M^2\left(\omega L_1-\dfrac{1}{\omega C_1}\right)}{r_1^2+\left(\omega L_1-\dfrac{1}{\omega C_1}\right)^2} \tag{3}$$

The apparent secondary inductance may be expressed as

$$\begin{aligned}L_a&=L_2-\Delta L_2\\&=L_2\left[1-\frac{\omega M^2}{L_2 z_1^2}x_1\right]\\&=L_2[1-K^2]\end{aligned} \tag{4}$$

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[1] "Electric Oscillations and Electric Waves," G. W. Pierce, p. 162.

in which K is a quantity which might be termed the "apparent coefficient of coupling between the circuits," and is defined as

$$K^2=\frac{\omega M^2}{L_2 z_1^2}x_1 \tag{5}$$

It will be noted that K^2 can be positive or negative, depending on the algebraic sign of x_1, whence it might appear that K would be imaginary when $\omega^2>\frac{1}{L_1C_1}$, or when x_1 is negative. This ambiguity arises from the arbitrary definition of K^2, and the algebraic sign of K should be considered as merely indicating the sense of the effects produced on the secondary. When $\omega^2>\frac{1}{L_1C_1}$, K^2 is positive, and $L_a>L_2$; and when K^2 is negative $\omega^2<\frac{1}{L_1C_1}$, and $L_a<L_2$.

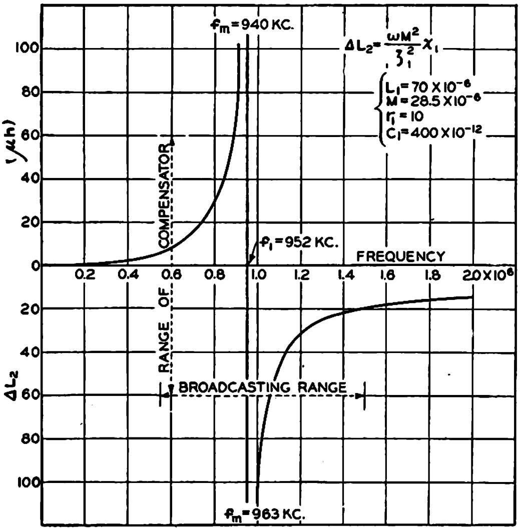

Fig. 2—Curves of Secondary Inductance Coil Variation.

The quantity K^2 has been so defined in order to be consistent with the definition of the ordinary coefficient of coupling, viz.,

$$k^2=\frac{M^2}{L_1L_2} \tag{6}$$

Thus, when $r_1=0$ and there is no condenser in the primary circuit ($C_1=\infty$) or, when the frequency is very much higher than the resonance frequency of the primary, expression (5) reduces to expression (6), or K^2 becomes equal to k^2. On account of this arbitrary definition, therefore, K can be given the same sign as K^2, the negative sign being disregarded when extracting the square-root.

From Eqs. (4) and (5) it follows that

$$\Delta L_2 = K^2 L_2 \tag{7}$$

K and ΔL_2 are zero when $\omega=0$ and when $\omega=\frac{1}{L_1C_1}$. At very high frequencies K approaches $k=\frac{M}{\sqrt{L_1L_2}}$ and ΔL_2 approaches $\frac{M^2}{L_1}$

The frequencies at which K and ΔL_2 are greatest are obtained by differentiating with respect to ω and equating to zero. A quadratic is obtained:

$$\left[\omega L_1-\frac{1}{\omega C_1}\right]^2\left[L_1+\frac{1}{\omega^2 C_1}\right]=\left[r_1{}^2+\left(\omega L_1-\frac{1}{\omega C_1}\right)^2\right]L_1$$

The solution gives

$$\omega_m{}^2=\frac{1\pm r_1\sqrt{\frac{C_1}{L_1}}}{L_1C_1-r_1{}^2C_1{}^2} \tag{8}$$

In practice $r_1{}^2C_1{}^2$ is generally $\ll L_1C_1$. Therefore

$$\omega_m{}^2=\pm\frac{r_1}{L_1}\omega_1-\omega_1{}^2$$

$$f_m{}^2=\pm\frac{r_1}{2\pi L_1}f_1-f_1{}^2 \text{ (approx.)} \tag{9}$$

Here again an ambiguity may arrive in respect to the algebraic sign of $\omega_m{}^2$. The sign of $\omega_m{}^2$, however, is again to be taken as merely signifying whether ω_m is > or < ω_1. As an example, for the values given in Fig. 3

$f_1=952{,}000$ cycles per second

$f_m=940{,}000$ and $963{,}000$ cycles per second.

A calculated curve showing the variation of ΔL_2 with frequency is shown in Fig. 2, for the values indicated.

Eqs. 2 and 4 suggest a means of easily measuring K. Suppose in Fig. 1 a voltage E at a given frequency is imposed upon the primary circuit, and C_2 is adjusted for maximum secondary current. Partial Resonance Relation S then obtains, and

$$C_2=\frac{1}{\omega^2 L_a}=\frac{1}{\omega^2 L_2(1-K^2)}$$

If now the primary circuit be broken, and the generator be inserted in the secondary circuit, the frequency being unchanged, a new setting of C_2 will be required to produce resonance, viz:

$$C_2'=\frac{1}{\omega^2 L_2}$$

The ratio of C_2 to C_2^1 results in

$$K^2=1-\frac{C_2'}{C_2}$$

As mentioned before, when there is no condenser in the primary this procedure gives the ordinary coefficient of coupling directly.

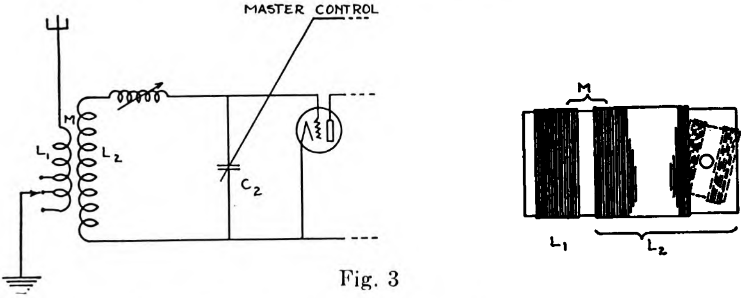

Fig. 3

The actual procedure employed is simple. The transformer is coupled loosely to an oscillator and a current indicator is placed in the secondary. Adjustment of C_2 for maximum indication of the meter furnishes C_2. The primary circuit is then opened and a readjustment of the condenser furnishes C_2'.

In radio receivers employing a master control, in which all of the variable tuning condensers are identical, in order that all the tuned circuits may be in resonance, it is necessary that all the secondary inductances be identical.

In all but the first stage, which is coupled to the antenna, the apparent secondary inductance does not vary appreciably with frequency in ordinary practice, on account of the high resistance in the primary circuits (the r_p of the tube to which it is connected), and it is practically equal to the inductance of the secondary alone.

The first stage, however, is subject to variation, due to the primary load. Such a variation for a circuit of the given constants is shown by Fig. 2. The apparent secondary inductance is greater or less than the true secondary inductance by the amount ΔL_2. In order to tune exactly, it should at all times have the same value as the apparent secondary inductance of the other stages of the receiver.

In order to compensate for this apparent variation of inductance, it must be possible to change L_2 in the same amount as ΔL_2 but in the opposite sense. By doing so, L_a will be kept constant, and the first stage will be in resonance as well as the other stages of the receiver.

Fig. 3 indicates how this may be done. When the field of the rotor, on which is wound part of the secondary winding, is perpendicular to the field of the stator, the total secondary inductance L_2 is the same as the secondary inductances in the other stages. On turning the rotor, L_2 may be increased or decreased in order to compensate for the change of secondary inductance, ΔL_2.

The range of such an experimental variometer was about 60 microhenries; its operating range is indicated on Fig. 2. The curves of Fig. 4 indicate the operation of the compensator when installed in a receiver, the values C_1 indicating the antenna capacity. Three taps were provided on the primary winding, numbered 1, 2, and 3 in Fig. 4. Tap No. 3 included the full winding, and tap No. 1 the smallest section. The heavy black lines in Fig. 4 indicate the tuning range of frequency for tap No. 3.

The curves are essentially flat excepting when tuning at frequencies at or close to the resonance frequency of the primary circuit. It is, therefore, necessary to rotate the variometer only a very slight amount while tuning from one frequency to another with the master control, excepting when operating with a large antenna on tap No. 3. The gaps in the black lines indicate regions of indefinite tuning.

It will be noted in Fig. 3 that the variable part of the secondary winding has been placed at the end of the coil away

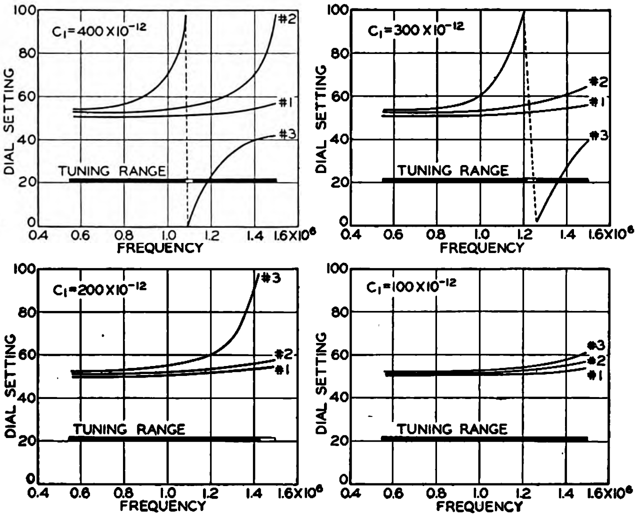

Fig. 4—Tuning Curves of Variometer.

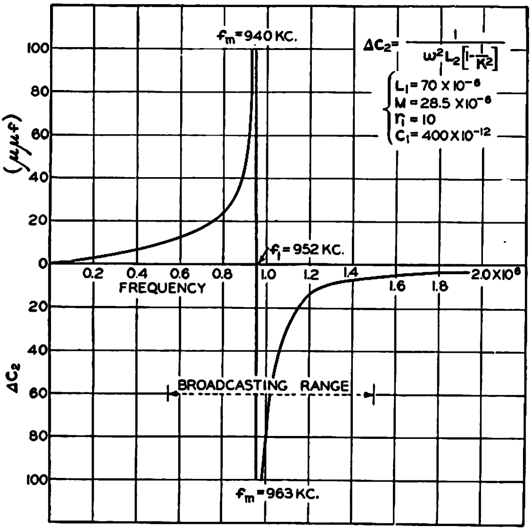

Fig. 5—Curves of Secondary Capacity Variation.

from the primary winding. This has been done to prevent, as far as possible, any increase of mutual inductance between the primary and secondary windings as the rotor is turned. An increase in the value of M as the rotor is turned to compensate for ΔL_2 would make ΔL_2 still greater and aggravate conditions, especially when operating near the resonance frequency of the primary. For the construction used M varied hardly more than a few microhenries, throughout the 180 degrees rotation of the variometer.

APPARENT CHANGE OF CAPACITY

If, instead of compensating for the effect of the primary by varying the inductance of the secondary, it is desired to do so by varying the capacity, (1) may be written

$$\omega L_2 = \frac{1}{\omega C_2} + \frac{\omega^2 M^2}{z_1^2} x_1$$

$$= \frac{1}{\omega \left[\dfrac{C_2}{1+\omega^2 L_2 C_2 K^2} \right]} = \frac{1}{\omega [C_2 + \Delta C_2]} = \frac{1}{\omega C_a} \tag{10}$$

Combining these equations and putting ΔL_2 for $L_2 K^2$ (Eq. 5) and $\frac{1}{L_a}$ for $\omega^2 C_2$;

$$\frac{\Delta C_2}{C_2} = -\frac{\Delta L_2}{L_2} \tag{11}$$

But $C_2 = \frac{1}{\omega^2 L_a}$. Therefore

$$\Delta C_2 = -\frac{1}{\omega^2 L_2 \left[\dfrac{L_2}{\Delta L_2} - 1 \right]} = +\frac{1}{\omega^2 L_2 \left[1 - \dfrac{1}{K^2} \right]} \tag{12}$$

Values of ΔC_2 are shown plotted in Fig. 5. The circuit arrangement is shown in Fig. 6.

A method of compensating for ΔC_2 is to arrange the tuning condenser of the input stage of the receiver so that while the rotor is turned by the same shaft that turns the rotors of the condensers in the other stages, the stator can be rotated by means of an individual control. The angle through which the stator must rotate is determined by the variation of ΔC_2 at and near the primary circuit resonance.

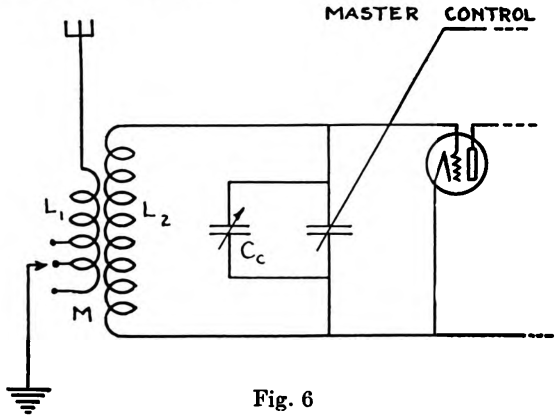

Fig. 6

A simpler means is to use a variable condenser operated by an individual control and connected in parallel with the main tuning condenser of the first stage. Since the compensating condenser is connected in parallel with the main tuning condenser of the first

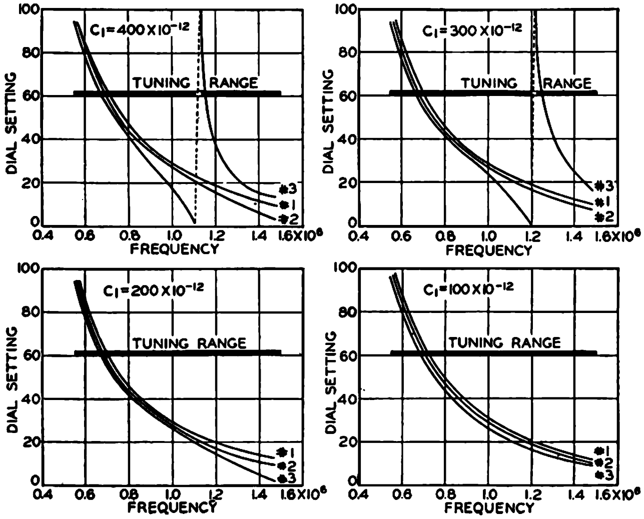

Fig. 7—Tuning Curves of Condenser.

stage, and C_a of this stage must be the same as the capacity in the other stages, the main tuning condenser of the first stage must be smaller than those of the other stages. For example, when the compensating condenser is set at its maximum, the total capacity must be the same as the maximum capacity in the other stages plus whatever is required to compensate for a given antenna load. Then the total capacity in the first stage at any frequency can be made greater or less than the capacity in the other stages at that frequency in an amount sufficient to compensate for the effect of the primary. The steepness of the curves of Fig. 7 is due to the fact that the capacity of the main condenser has been decreased, and consequently the additional condenser performs the two functions of assisting the main condenser in tuning and of compensating for the antenna load simultaneously This, however, causes no difficulty in tuning, although of course the flatter curves obtained with the variometer and the rotating stator are more desirable.

For this case let

C_2 = capacity of the tuning condenser in the second or third stage of the receiver.

mC_2 = capacity of main tuning condenser of the first stage

C_c = capacity of compensating condenser

Then

$$C_c = C_2(1-m) + C_2 \tag{13}$$

$$= \frac{1}{\omega^2 L_2\left[\dfrac{1}{1-m}\right]} + \frac{1}{\omega^2 L_2\left[1-\dfrac{1}{K^2}\right]} \tag{14}$$

The first term of the right hand member of (14) is the difference of capacity between C_2 of the other stages and the capacity of the main tuning condenser of the first stage. The second term is the capacity required to compensate for the antenna load. Operating curves for such an arrangement are shown in Fig. 7. the heavy black lines indicating the tuning range for tap No. 3 and the gaps in these lines indicating the regions of indefinite tuning.

THE CAUSE AND PREVENTION OF HUM IN RECEIVING TUBES EMPLOYING ALTERNATING CURRENT DIRECT ON THE FILAMENT*

By

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Summary—*The various factors affecting the variation of plate current of a receiving tube when the filament is heated with a.c. are analyzed, and their composite effect considered. The variation in temperature of the filament is found to be of minor importance while the predominant factor is the voltage drop along the cathode. The smallest variation in I_p is obtained from a tube having a properly designed V-shaped oxide filament. When straight filaments are used, their characteristics should be so chosen that the magnetic and electrostatic effects are neutralized.*

THE direct heating of the cathode of a thermionic tube by alternating instead of direct current is generally attended by a variation in plate current having a fundamental frequency which is twice that of the a.c. supply. When a so-called a.c. tube is used for the reception of radio signals, this variation in plate current gives rise to an objectionable hum in the sound output device. The design of a tube for this service necessarily entails a unique filament construction in which the main consideration is to minimize the influence of the factors tending to produce the hum. The purpose of this paper is to segregate these factors and to indicate in a qualitative manner their importance in affecting the magnitude of the composite hum. A rigorous treatment of the problem has not been attempted, it being desired to show only the order of magnitude of the quantities involved.

A careful analysis has revealed the following possible sources of hum:

(1) Temperature variation of the filament due to the sinusoidal power supply resulting in a variation in plate current.

(2) Variation in plate current due to the voltage drop along the filament.

(3) Variation of plate current due to the effect of the magnetic field of the alternating current.

In addition, there are possibilities of hum resulting from an emf induced in the metallic parts of the tube structure by the

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filament heating current. However, in dealing with this detail it was impossible to detect any effect originating from this source in a.c. tubes so designed as to obtain the optimum relation among the three previously mentioned factors such as in the Radiotron UX-226. Promiscuously arranged wiring in the associated tube circuits may also result in induced currents which increase the hum.

In the treatment of the main causes of hum, each has been taken up separately and considered as though it alone produces the hum, the effect of each on the resultant hum then being considered.

TEMPERATURE EFFECT

It is evident that a filament supplied by a.c. power will undergo a temperature variation about a mean value. Whether or not this temperature change is effective in varying the electron emission from the filament depends upon the magnitude of the temperature change and upon the proximity of the magnitude of the electron current to its saturation value. The general impression has prevailed that most of the hum from a.c. tubes is due to temperature effects. The following considerations indicate that the temperature change which filaments of even quite small cross-section undergo is very small. To measure directly this change in temperature is difficult. It is possible, however, to calculate the theoretical value by making assumptions which will cause the calculated value, if in error, to be larger than the true value.

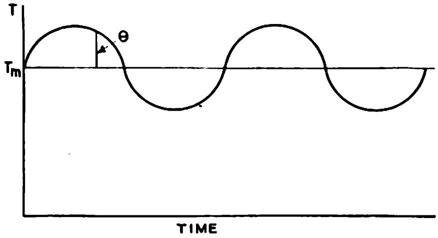

Fig. 1

Consider a length of filament from which the heat dissipated (by radiation and conduction) is expressed as $f(T)$. $f(T)$ is assumed to remain constant over the temperature range through which the filament passes. Although this is not strictly true, since the rate of radiation varies with the temperature, the variations of T will be found to be of small magnitude and symmetrical about a mean value so that the error introduced is quite small. Equating the rate at which energy is received by the filament to the rate at which it is dissipated and stored in the filament:[1]

$$RI^2 \sin^2 \omega t = f(T_m + \theta) + c\frac{d}{dt}(T_m + \theta) \tag{1}$$

where T_m = Average temperature of the filament (Fig. 1)

θ = Temperature above or below T_m at any time t.

R = Resistance of the section of the filament having heat capacity c. Both c and R are considered to be constant over the range $T_m \pm \theta$ which will be true for most cathode materials for small values of θ only.

$I \sin \omega t$ = Filament heating current.

Expanding $f(T_m+\theta)$ into a Taylor's series and neglecting second and higher powers of θ

$$RI^2 \sin^2 \omega t = f(T_m) + \theta f'(T_m) + c\frac{dT_m}{dt} + c\frac{d\theta}{dt} \tag{2}$$

Since $f(T_m)$ is the rate of heat dissipation at T_m:

$$f(T_m) = \frac{RI^2}{2}$$

Also since $f(T_m)$ is considered as being constant,

$$f'(T_m) = 0$$

Also

$$c\frac{dT_m}{dt} = 0$$

Equation (2) then reduced to:

$$\frac{RI^2}{2}(1 - \cos 2\omega t) = \frac{RI^2}{2} + c\frac{d\theta}{dt}$$

$$\frac{d\theta}{dt} = -\frac{RI^2}{2c} \cos 2\omega t$$

$$\theta = \frac{RI^2}{4c\omega} \cos\left(2\omega t - \frac{\pi}{2}\right) \tag{3}$$

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[1] Bockstahler—*Physical Review*, May, 1925.

This indicates that θ varies with twice the frequency of the heating current and lags by an angle $\pi/2$.

The maximum value of θ is given by:

$$\theta_{\max}=\frac{RI^2}{4c\omega} \tag{4}$$

Modifying (4) so that it can be applied more readily, one obtains:

For a filament of circular cross section:

$$\theta_M=\frac{0.152W}{s\rho df} \tag{5}$$

where θ_M = Range in temperature in deg. C from $+\theta_{\max}$ to $-\theta_{\max}$.

W = Heat dissipated from unit area of the filament expressed in watts per cm^2

s = Specific heat of the filament material

ρ = Density of the filament material in grams per cm^3

d = Diameter of the filament in centimeters

f = Frequency of the a.c. power supply

For rectangular cross section:

$$\theta_M=\frac{0.076W(a+b)}{s\rho abf} \tag{6}$$

where a and b are the length and width of the cross section in centimeters. A circular cross section, of course, results in the lowest value of θ_M for a given cross-sectional area.

It will be noticed that no correction has been applied to compensate for the change in heat capacity effected by the oxide coating on oxide filaments. This error is undoubtedly appreciable and tends to make the values calculated by the above formulas larger than the true value. However, as indicated previously, the formulas are not exact but give an indication of the maximum possible temperature excursions to which an a.c. heated filament is subjected. It will be evident from the following calculated values of θ_M for several well-known types of tubes, that the temperature variation is so small that negligible variation in emission can be attributed to this cause.

Tube	Type of Filament	θ_M	
UX-226	Oxide	1 deg. C	
UX-112	"	3 "	f=60 cycles
WX-12	"	10 "	
UX-201A	Thor. Tungsten	16 "	

A consideration of the familiar temperature saturation curve (Fig. 2) for a two-electrode thermionic tube will show that the cathode must undergo quite large temperature variations before sub-saturation values of plate-filament current are affected. Curve OAB represents variation in plate current with variation in plate-filament potential for a filament temperature T_1. Increasing the cathode temperature from T_1 to T_2 raises the upper bend of the curve from A to C, but has no effect upon the value of I_p corresponding to a value of E_p such as E_p' which is considerably smaller than the saturation value of E_p. In actual three-electrode tubes used for receiving purposes, the filaments are capable of giving saturation values of I_p of from 25 to 200 milliamperes, while I_p under normal operating conditions seldom exceeds from 2 to 10 milliamperes. It is more or less evident then that an appreciable cathode temperature variation can be tolerated without producing any change in the operating value of I_p.

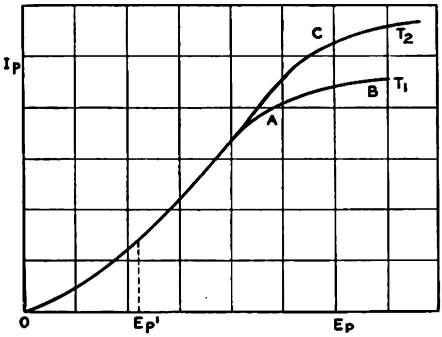

Fig. 2

VARIATION IN PLATE CURRENT DUE TO THE VOLTAGE DROP ALONG THE CATHODE

The grid and plate returns to the cathode of an a.c. operated tube are connected to the center rather than to one side of the filament, as is usually the case with receiving tubes for d.c. service. This center return is effected either by a tap on the transformer filament winding or by means of a voltage divider (commonly known as a potentiometer in radio parlance).

It is desirable to determine the effect produced upon the plate current by having the return to the filament connected to some point other than the electrical center. Consequently, in considering the variation in plate current produced by the voltage drop along the filament, a cylindrical two-element structure in which the anode return is connected to the cathode at a point distant mL from one end, L being the length of the structure considered and $O<m<1$, will be taken up. The cathode is considered to be a cylinder of small radius coincident with the axis of the cylinder forming the anode, the entire structure being a section taken from an infinite length. The cathode, moreover, will be considered to give uniform emission throughout its length under equi-potential and equi-temperature conditions. Also the initial electronic velocities are assumed to be zero.

Langmuir's relation[2]

$$i=\frac{2}{9}\sqrt{\frac{e}{m}}\,\frac{L}{r}E^{3/2}$$

is used in the form

$$I_p=kLE_p{}^{3/2}$$

Considering the potential of the point of the filament to which the plate return is connected (Fig. 3) to be zero, the potential difference between the plate and point A of the filament will e^2

$$E_p-\frac{x}{L}E_f\sin t(E_p\gtreqless E_f)$$

The emission from length dx of the cathode will be:

$$\Delta I_p=k\left(E_p-\frac{x}{L}E_f\sin\omega t\right)^{3/2}dx$$

From the entire length L of the cathode:

$$I_p=k\int_{-mL}^{(1-m)L}\left(E_p-\frac{x}{L}E_f\sin\omega t\right)^{3/2}dx \text{ where } O<m<1$$

$$=-\frac{2kE_p{}^{5/2}L}{5E_f\sin\omega t}\left[\left(1-\overline{1-m}\frac{E_f}{E_p}\sin\omega t\right)^{5/2}\right.$$

$$\left.-\left(1+m\frac{E_f}{E_p}\sin\omega t\right)^{5/2}\right]$$

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[2] Van der Bijl, "Thermionic Vacuum Tube."

$$= kLE_p{}^{3/2}\left\{1-\frac{3}{4}\,\frac{E_f}{E_p}\sin\omega t\left[(1-m)^2-m^2\right]+\frac{1}{8}\left(\frac{E_f}{E_p}\sin\omega t\right)^2\right.$$

$$\left.\left[(1-m)^3+m^3\right]+1/64\left(\frac{E_f}{E_p}\sin\omega t\right)^3\left[(1-m)^4-m^4\right]\cdot\cdot\cdot\right\} \tag{7}$$

For $m=0.5$ which is the case when connection is made to the electrical center of the filament:

$$I_p=KLE_p{}^{3/2}\left\{1+1/32\left(\frac{E_f}{E_p}\sin\omega t\right)^2\right.$$

$$\left.+\frac{3}{10240}\left(\frac{E_f}{E_p}\sin\omega t\right)^4\cdot\cdot\cdot\right\} \tag{8}$$

For a value of $E_f/E_p=10^{-1}$ the amplitude of the third term in the above series (8) is 10^{-4} times the amplitude of the second term,

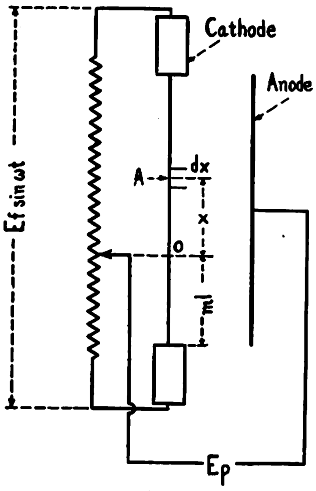

Fig. 3

enabling one to express the plate current to a first approximation as

$$I_p= kLE_p{}^{3/2}\left[1+\frac{1}{32}\left(\frac{E_f}{E_p}\right)^2\sin^2\omega t\right] \tag{9}$$

$$= kLE_p{}^{3/2}\left[1+\frac{1}{64}\left(\frac{E_f}{E_p}\right)^2-\frac{1}{64}\left(\frac{E_f}{E_p}\right)^2\cos 2\omega t\right] \tag{10}$$

Equation (10) indicates that the plate current has added to it a d.c. value given by the second term of the equation and an a.c. component of twice the frequency of the power supply given by the third term.

It is also desirable to determine the value of plate potential which, if added to E_p when the cathode is emitting under equipotential conditions, would give the same expression for I_p as (10). Let this value of potential difference be represented by ϕ. Then from (9)

$$I_p = kL(E_p+\phi)^{3/2} = kLE_p^{3/2}\left[1+\frac{1}{32}\left(\frac{E_f}{E_p}\right)^2 \sin^2 \omega t\right]$$

$$(E_p+\phi)^{3/2} = E_p^{3/2}\left[1+\frac{1}{32}\,\frac{E_f^2}{E_p^2}\sin^2 \omega t\right]$$

To a first approximation, omitting terms of higher order than the second

$$\phi = \frac{1}{48}\,\frac{E_f^2}{E_p}\sin^2 \omega t \qquad (11)$$

$$= \frac{1}{96}\,\frac{E_f^2}{E_p}(1-\cos 2\omega t) \qquad (12)$$

Before discussing the application of these formulas to three-element tubes, the physical significance of the preceding development will be considered. In Fig. 4a the effect of voltage drop in the filament is graphically indicated. AB represents the filament, one end of which is at the maximum positive potential and the other end maximum negative potential, the center O being at zero potential. Ordinates represent emission, and are greatly exaggerated to make the phenomena more evident. CD shows how the emission varies over the length of the filament when A is +, that is, the difference in potential between the anode and A is $E_p - E_f/2$ and between the anode and B is $E_p + E_f/2$. This emission variation is a three-halves power relation with respect to E_p.

If the emission were a linear relation, conditions would be represented by Fig. 4b. Line EF (as in 4a) indicates the value of the emission at the instant the potential difference between A and B is zero, resulting in a potential difference from every point on the filament to the plate of E_p. When the potential between A and B is other than zero, the average emission from the entire filament remains the same regardless of the magnitude of the drop between *A* and *B*, since the increase in emission from one half of the filament represented by *FO'D* equals the decrease in emission from the other half, represented by *CO'E*. If it were possible to design a tube having a linear relation between cathode emission and anode-cathode potential, supplying the filament with a.c. would produce no hum due to potential drop along the cathode. In this case the emission center would always be at the same point as the electrical center, the emission center of the cathode being defined as that point which divides the cathode into two equi-emitting parts, so that each part of the cathode so divided contributes half of the emission forming the plate current.

However, when dealing with actual conditions where the emission from the cathode is governed by the three-halves power

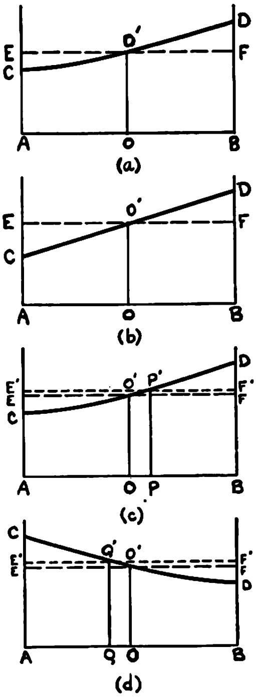

Fig. 4

relation, supplying the cathode with a.c. power causes the emission center to oscillate about instead of remaining coincident with the electrical center. In Fig. 4c, the line *E'F'* is so drawn that the area $PP'DB$ equals the area $PP'AC$, making P the emission center at the instant when the potential difference between A and B causes the emission to be represented by the line CD. Fig. 4d represents conditions when the potentials of A and B are of opposite sign to those indicated for Fig. 4c. The emission center has moved along the filament from P to Q on the other side of the electrical center. For intermediate values of potential drop along the cathode, the emission center assumes some corresponding point between P and Q.

It is evident that a difference in potential exists between the electrical and emission centers of the cathode and that the electrical center of the cathode is not at zero potential with respect to the cathode but has impressed upon it a sinusoidal potential. To determine the approximate magnitude of this potential, consider the familiar expression

$$I_p = f(kE_g + E_p)$$

which specifies that in a three-electrode tube the electrostatic field effective in controlling the electron emission from the cathode is expressed as $(kE_g + E_p)$, where (k) is a structural constant of the tube and is a measure of the shielding action of the grid. It is seen that the electric field in the two-element tube is determined by E_p alone. Adapting equation (12) to a three-electrode structure, one obtains

$$\phi' = \frac{E_f^2}{96(kE_g + E_p)}(1 - \cos 2\omega t) \qquad (13)$$

ϕ' acts in the grid circuit as well as in that of the plate, since the return lead from the grid to the cathode is also connected to the electrical center of the cathode.

The effect, then, of heating the filament of a three-electrode tube by alternating current is that of introducing a potential in the plate circuit of the tube given by

$$(\mu + 1)\phi' = \frac{(\mu + 1)E_f^2}{96(kE_g + E_p)}(1 - \cos 2\omega t) \qquad (14)$$

where μ is the dynamic amplification factor of the tube. In actual tubes, where the theoretical conditions are considerably modified, the magnitude of the above quantities will be at variance with the true values. However, they serve to indicate the nature of the phenomena occurring and afford a basis for so choosing tube dimensions and characteristics as to decrease the variation in I_p to a minimum.

VARIATION OF PLATE CURRENT DUE TO THE EFFECT OF MAGNETIC FIELD OF THE ALTERNATING CURRENT

The influence of the magnetic field due to the heating current of the filament is in general a negligible quantity when dealing with currents of the order of magnitude encountered in the filaments of receiving tubes. Due to the fact, however, that the variation in I_p produced by a.c. on the filament is a second order quantity itself, one must take the effect of the magnetic field into consideration. Referring to Fig. 5, AA is the cylindrical anode of a two-element thermionic tube and F is the filament coincident with the axis of the cylinder. OD represents the path followed by an electron emitted perpendicular to the filament through which the filament current is zero. Under these conditions there will be no magnetic field and the velocity with which the electron strikes the plate will be determined wholly by the difference in potential between the anode and cathode and upon the initial velocity of the electron. OB represents the path followed when the field due to the heating current is of just sufficient intensity to cause the electron to return to the filament before it encounters the plate. A. W. Hull[3] has given equations for the behavior of an electron under such conditions. OC indicates the path followed when the magnetic field assumes a value intermediate between zero and the critical value necessary for path OB.

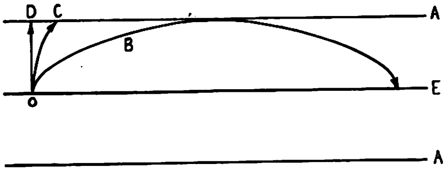

Fig. 5

It is evident that the radial velocity of the electron under conditions attending the pursuing of path OC will be different than that for path OD when considering points equi-distant from the cathode. This is due to the unvarying magnetic field being unable to contribute any energy to the electron, it being assumed that the filament is heated by direct current or low-frequency alternating current. The resultant velocity then at any distance from the cathode will be the same for path *OD* as for *OC*.

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[3] A. W. Hull, *Physical Review*, May, 1925.

The equation

$$\begin{aligned} I_p &= 2\pi R d v \\ &= K v \end{aligned} \tag{15}$$

represents the current in the external circuit per unit length of the two-element structure under consideration where d is the electron density at the plate, v is the radial electron velocity at the plate, and R the radius of the plate.

To obtain an expression for the hum or I_p variation associated with a tube as a result of the varying magnetic field about the filament when an a.c. power supply is used, the difference in the velocity with which an electron strikes the plate in the presence of a magnetic field and with no magnetic field present may be compared. However, for the present development it was found unnecessary to obtain a quantitative expression for this effect. It is sufficient to note that v and consequently I_p vary sinusoidally with and at twice the frequency of I_f. If the filament voltage and current are in phase, the hum component due to the effect of the magnetic field is 180 deg. out of phase with that due to the filament potential drop. This is evident from a consideration of the conditions existing when there is maximum potential difference between the ends of the filament. At this instant the voltage drop along the filament tends to produce maximum emission to the plate, while at the same time the maximum current through the filament establishes the maximum value of magnetic field which tends to reduce the emission to the plate in the manner previously indicated.

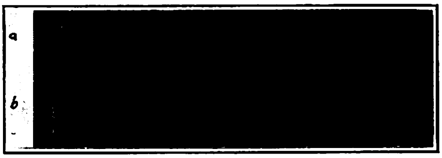

Fig. 6

APPLICATION OF THE PRECEDING CONSIDERATIONS

A considerable number of three-electrode tubes were constructed, having different filament characteristics, in an attempt to obtain a design which would be satisfactory for use as an amplifying tube in radio receiving sets. Oxide-coated filaments were used exclusively and it was found possible to use as small a diameter filament as 0.103 millimeters and still obtain very low hum values. To justify the assumption that the temperature variation of this size filament (about 0.9 deg. C for 60 cycles by calculation) is so small as to have no effect upon varying the emission, an oscillogram of the plate current of a tube having a filament whose calculated temperature variation was about 1.2 deg. C was taken, and is shown by trace (b) of Fig. 6. In Fig. 7 is given the wiring diagram of the apparatus used to obtain this oscillogram. The double-frequency curve was traced by oscillograph element No. 1 which was actuated by the current in the secondary winding of a step-down transformer whose primary winding was in the plate circuit of the tube under consideration. The lower frequency curve represents the filament voltage applied to the tube and was traced by element No. 2 which was actuated by a similar step-down transformer having the same d.c. component flowing through its primary.

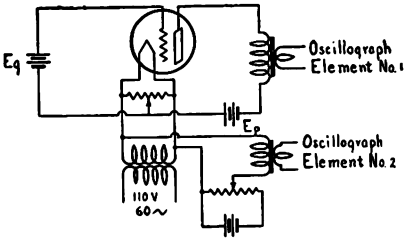

Fig. 7

It is evident that the I_p variation (hum) is a double-frequency effect (compared to the frequency of the filament heating current) and is very nearly in phase with the filament voltage. These curves are very similar to the plot (Fig. 12) of $A/2\,(1-\cos 2\omega t)$ and $A\sin\omega t$ predicted by equation (10) for the hum.

Trace (a) of Fig. 6 was taken under conditions similar to those for trace (b) except that a tube having a thoriated tungsten filament for which the calculated temperature variation was approximately 15 deg. C was substituted for the oxide-filament tube. The filament voltage drop and current were the same for both of the tubes, but the mutual conductance of the oxide filament tube was about three times that of the other accounting for the greater amplitude of the hum curve given by trace (b).

The significant fact indicated by trace (a) is that the hum is out of phase with the filament voltage. Equation (3) shows that the temperature change lags behind the filament power supply so that if part of the hum is due to temperature effects, the resultant hum should be out of phase with the filament voltage by an angle whose magnitude would depend upon the relative magnitudes of the hum components.

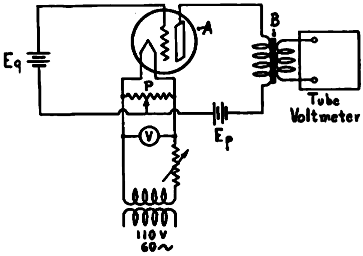

Fig. 8

In studying the other causes affecting the hum, it was considered permissible to assume that the hum due to temperature variation was practically zero since the calculated temperature change for the filaments used did not exceed about 1.2 deg. C.

Fig. 8 is the wiring diagram for the apparatus used to measure the intensity of the hum produced by the tubes studied. The tube under observation is indicated by *A*. The filament in all cases was supplied by 60-cycle current from a step-down transformer. The plate and grid returns were made to the variable contact of a rheostat *P* connected across the filament terminals. The plate, grid, and filament voltages could be adjusted to any desired value. A vacuum-tube voltmeter connected to the secondary winding of a nickel-iron core audio-frequency amplifying transformer *B* whose primary winding was in the plate circuit of the tube being measured indicated the magnitude of the hum. This voltmeter was calibrated in terms of millivolts at 120 cycles applied to the primary winding of transformer *B*.

Equation (13) indicates that a definite a.c. voltage is applied to the grid of a tube whose filament is heated by a.c. Increasing the amplification factor of the tube should consequently increase the hum obtained from the tube. To determine the authenticity of this deduction, a number of tubes were constructed as nearly alike as possible except for a difference in grid mesh so as to obtain tubes of similar filament construction but with different amplification factors. Fig. 10 shows the variation of hum with amplification factor as determined by measurements of these tubes. Equation (13) predicts a linear relation between the hum and the amplification factor. However, the tubes which had the higher amplification factors necessarily had high impedances. The transformer used in the plate circuit of the a.c. tube in the hum measuring set had too low a primary impedance to give as great amplification for the higher impedance tubes as for the lower impedance tubes, accounting in part for the bending over of the curve.

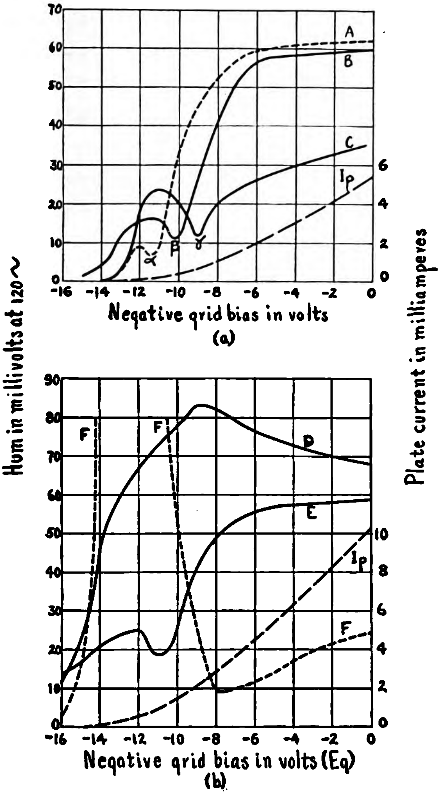

Fig. 9—Variation of Hum with E_g.
$E_p = 90$ volts; $\mu = 8.0$

In Fig. 9 are given curves for the variation of hum with grid bias as measured by the hum measuring set, for several tubes whose filament characteristics are given in Table I.

TABLE I

TUBE	TYPE OF FILAMENT	I_f(R.M.S.)	E_f(R.M.S.
A	0.170 mm. Round Oxide Single Strand	1.2 amps.	0.8
B	0.155 " " " " "	0.9 "	0.7
C	0.103 " " " " "	0.5 "	1.1
D	0.178 (Approx)" " Two Strands	2.5 "	1.0
E	0.155 " " " " "	1.8 "	1.7
F	(UX-226) Ribbon Oxide V-Shape	1.05 "	1.5

All of the tubes had a flat anode structure and the tubes whose curves are given in (9a) had filaments consisting of a single strand of the oxide-coated base metal through the center. The minimum hum points, (α, β, γ) give evidence of the balancing effect of the two hum components and as the diameter of the filament is decreased, the minimum position moves toward more positive values of grid bias. Further decreasing the filament diameter should accordingly move the minimum hum position farther to the right. It was considered, however, that filaments of smaller diameter than that of tube *C* (0.103 mm) would probably be subject to objectionable temperature variations and would also be very fragile. In order to cause the minimum position to occur on the operating part of the characteristic ($E_g = -4$ to -6 volts for $E_p = 90$ volts), a different filament construction had to be resorted to.

This was effected by use of a *V*-shaped filament. Curve *F* of Fig. 9b gives the variation of hum with E_g for the Radiotron UX-226 which incorporates this type of filament. It will be observed that the hum produced by this tube at the operating value of E_g is considerably lower than that obtained from any straight filament construction considered, regardless of the magnitude of the voltage drop along the filament.

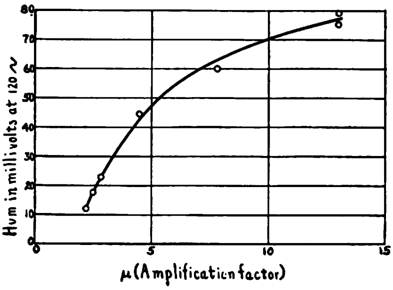

Fig. 10—Variation of Hum with Amplification Factor.

The explanation advanced for the low value of hum obtained with this tube is that there is a passage of electrons between the two legs of the *V*-shaped filament. Considering the instant when there is maximum potential difference between the ends of the filament, which is in a negative or only slightly positive electrostatic field due to the negative potential on the grid, some of the electrons emitted by the negative end of the filament are attracted to the positive end instead of to the plate. It is at this instant that there is maximum emission from the filament due to the maximum potential drop along the filament. This increase in emission is partly reduced by the exchange of electrons between the two portions of the filament.

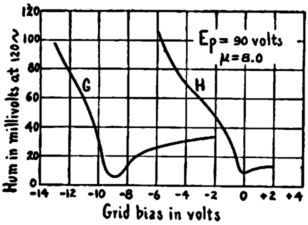

Fig. 11—Hum Curves for V-Shaped Filament Tubes.

Fig. 11 shows the effect produced on the hum curve by varying the distance between the ends of the filament *V*. If the assumption that the reduction in hum results from a flow of electrons between the filament legs is correct, bringing the ends of the *V* closer together should cause a greater number of electrons to pass to the positive side of the *V* due to the increased potential gradient. The minimum hum position should then move to a more positive value of E_g where I_p is greater. Curve *G* represents the variation of hum with E_g for the greatest spread between the ends of the filament possible for the tube structure used. Curve *H* shows the variation of hum when the sides of the *V* are almost in contact with each other. The other structural details of these two tubes were as nearly similar as possible, and the same filament and plate voltages were used in making the measurements.

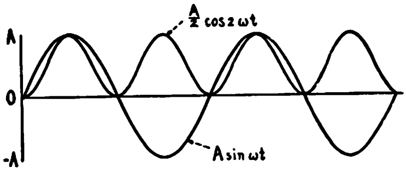

Fig. 12

CONCLUSION

(1) The foregoing considerations have shown that the temperature variations encountered when oxide filaments of ordinary size heated by a.c. power are used are so small that it is unnecessary to resort to a circular cross section and heavy current construction.

(2) Neutralizing Effects:

a. The hum component due to temperature variation cannot be neutralized by the other factors since it lags by very nearly 90 deg. behind the power supply. This component, however, can be made so small as to be negligible.

b. With the use of straight filaments, the lowest hum is obtained by so choosing the filament characteristics that the hum components due to voltage drop and magnetic field neutralize each other.

c. The use of *V*-shaped filaments enables a very low minimum hum value to be obtained. The position of minimum hum on the E_g axis can be varied at will by choosing the proper filament design.

(3) The effect of the alternating voltage drop along the filament is to place a sinusoidal potential on the grid. Consequently, the hum produced by an a.c. tube increases with the amplification factor.

THE INTERNATIONAL UNION OF SCIENTIFIC RADIOTELEGRAPHY*

By

J. H. Dellinger
(Chief of Radio Section, Bureau of Standards, Washington, D.C.)

THE International Union of Scientific Radiotelegraphy was organized in 1919. It is commonly designated as the URSI (from its French name, Union Internationale de Radio-télégraphie Scientifique). Its aims are: (1) to promote the scientific study of radio communication; (2) to aid and organize researches requiring cooperation on an international scale and to encourage the discussion and publication of the results; (3) to facilitate agreement upon common methods of measurement and the standardization of measuring instruments. The International Union itself is an organization framework for carrying on the international phases of the administrative work, and operates under the International Research Council. The actual technical work is done by the various national sections. The officers are: President, General G. Ferrié, France; Vice-Presidents, Dr. L. W. Austin, United States, Dr. W. H. Eccles, England, Dr. G. Vanni, Italy, Dr. V. Bjerknes, Norway; General Secretary, Dr. R. B. Goldschmidt, Belgium.

The General Secretary's office, and the headquarters of the organization, are located at 54 Avenue des Arts, Brussels, Belgium. The organization is financed by contributions of the various governments, through the International Research Council, but the total amount of such support is very small. In fact, the routine of the organization is carried on largely through voluntary service. The General Secretary's office carries on the international correspondence and duplicates documents originating in the various National Sections, distributing them to all the other Sections.

The American Section was organized in 1920, under the National Research Council. It is made up of a group of technical committees and an executive committee. The executive committee is as follows:

Dr. L. W. Austin, chairman, representing the International Union as its vice president.

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* Original Manuscript Received by the Institute, May 26, 1928. Publication Approved by the Director of the Bureau of Standards in the U. S. Department of Commerce.

Dr. J. H. Dellinger, technical secretary, representing the Department of Commerce.

Dr. W. E. Tisdale, corresponding secretary, representing the Division of Physical Sciences, ex-officio.

Prof. D. C. Miller, representing the Division of Physical Sciences, ex-officio.

Dr. A. N. Goldsmith, representing the Institute of Radio Engineers.

Major-General G. S. Gibbs, representing the Army, ex-officio.

Major W. R. Blair, representing the Army.

Captain T. T. Craven, representing the Navy, ex-officio.

Dr. A. H. Taylor, representing the Navy.

Members at large: E. F. W. Alexanderson, Dr. A. E. Kennelly, Major-General G. O. Squier, Dr. Wm. Wilson, Prof. E. M. Terry, F. Conrad, G. W. Pickard, Prof. E. L. Chaffee, Dr. G. Breit, H. T. Friis.

The American section holds an annual meeting in Washington, D.C., in April. At this meeting the chairmen of the several technical committees present progress reports on the status of work in the field of the committee, and technical papers are presented by the various members of the committees.

Membership in the American Section is open to anyone actively interested in radio. Anyone desiring to join should write to the Technical Secretary, stating in which of the technical committees he is most interested. The technical committees are as follows:

1. Radio measurements.
1a. Measurement of interference.
2. Wave propagation.
2a. Wave direction.
2b. Phenomena above 3 megacycles.
3. Atmospherics.
4. Cooperation.
5. Oscillations.

The chairman of the first committee is Dr. J. H. Dellinger, Chief of the Radio Section of the Bureau of Standards. This committee is concerned with radio measurements and standards, particularly as regards measurement work performed by other committees of this organization. The committee promotes and participates in the improvement and standardization of methods for measuring frequency, field intensity of received waves, intensity of atmospheric disturbances, automatic recording devices, fading, and other types of radio measurements.

The committee on the measurement of interference is headed by E. F. W. Alexanderson, of the General Electric Company. It studies all forms of interfering signals and radiation, other than atmospheric disturbances. It deals particularly with undesired emissions, including harmonics and broad waves.

The committee on radio wave propagation is headed by Dr. L. W. Austin, in charge of the Bureau of Standards laboratory for special radio transmission research. This committee encourages and conducts measurements upon received field intensities. It investigates, in connection with these measurements, diurnal and seasonal variations and effects of transmission under day and night conditions, sunset and seasonal effects, and correlations with other natural phenomena.

The chairman of the committee on radio wave direction is Dr. Gregory Breit, department of terrestrial magnetism, Carnegie Institution. This committee encourages and correlates research on the variations of wave direction with time, effects of topography on direction-finder work, and the bearing of polarization and other variations upon apparent observed phenomena.

The committee on wave phenomena above three megacycles is under the direction of Dr. A. H. Taylor, of the Naval Research Laboratory. The work covers such phenomena as daily and seasonal variations of received field intensities from high-frequency stations, character of fading and atmospherics determination of skip distances, comparison of radio phenomena with magnetic and solar variations, weather, etc.

The committee on atmospheric disturbances, under H. T. Friis, of the Bell Telephone Laboratories, as chairman, promotes and coordinates measurements upon the intensity and direction of atmospherics, including daily and seasonal variations, characteristics of the various types of atmospherics, their frequency distribution, and methods of recording and measuring.

Professor A. E. Kennelly, of Massachusetts Institute of Technology, is chairman of the committee on cooperation. This committee seeks to enlist the cooperation of experimenters, amateurs, and others in radio observation programs. It also maintains contact with the workers in related sciences, such as geophysics, astronophysics, terrestrial magnetism, and atmospheric electricity.

The chairman of the committee on "oscillations" is Professor E. L. Chaffee of Harvard University. This committee promotes progress in the general theory of oscillations and its radio applications. Its interest covers, e.g., the general theory of vacuum tubes, telephone receivers, modulation, and circuit networks. It covers in general the broad radio field not included under the other committees.

The International Union itself holds annual meetings. The 1927 meeting was held in October in Washington, at the same time as the International Radio Conference. The 1928 meeting is to be held in September in Brussels, Belgium. The Washington meeting was prolific in the presentation of important scientific results and in the exchange of information and ideas among the radio scientists of the world. It was attended by about 70 persons from the following countries: Australia, Belgium, Canada, France, Germany, Great Britain, Holland, India, Ireland, Italy, Japan, Norway, Switzerland, and the United States. Sixteen sessions were held, between October 10 and 28. These included one public and two general sessions, held under the chairmanship of the President, General Ferrié, and sessions of the several "commissions." The presiding officers at the latter were as follows:

I. Radio Measurements, Dr. D. W. Dye (England)
II. Wave Phenomena, Dr. L. W. Austin (U. S. A.)
III. Atmospherics, Prof. R. Mesny (France).
IV. Cooperation, Prof. G. Vanni (Italy).

The great variety of work covered is illustrated by the titles of papers presented at the public session such as:

Employment in combination of photoelectric cells and vacuum tubes in the solution of various problems of time measurements. General G. Ferrié.

International comparison of frequency standards. Dr. J. H. Dellinger.

Precision determination of frequency. J. W. Horton and W. A. Marrison.

The Navy's primary frequency standard. R. H. Worrall and R. B. Owens.

A radio-frequency oscillator for receiver investigations. G. Rodwin and T. A. Smith.

An automatic recorder for radio signals and atmospheric disturbances. E. B. Judson.

Experiences in radio compass calibration. F. A. Kolster.

Apparent night variations in crossed-coil radio beacons. H. Pratt.

Ionization in the outer atmosphere of the earth. Dr. E. O. Hulburt.

The constitution of the earth's atmosphere up to great heights. H. B. Maris.

Ultrashort waves. Prof. R. Mesny.

Height of the reflecting layer in August, 1927, and the effect of the disturbances of August 19. O. Dahl and L. C. Gebhardt.

The relation between radio reception, sun-spot position, and area. G. W. Pickard.

Solar activity and radio transmission. Dr. L. W. Austin and Miss I. J. Wymore.

The effect of polarization on the received signal strength. Dr. B. van der Pol.

Investigation of downcoming waves. Prof. E. V. Appleton.

Experiments on radio wave projectors. E. F. W. Alexanderson.

Seasonal variation of the 20-meter wave from Nauen in Japan. T. Nakagami and T. Ono.

Diurnal variation in signal strengths of short waves. T. Nakagami and T. Ono.

Note on short-wave long-distance transmission. T. Minohara and K. Tani.

Directional observations on atmospherics in Japan. E. Yokoyama and T. Nakai.

Relations between atmospherics and meteorological phenomena. R. Bureau.

The office of the general secretary is preparing a special publication which will include all the papers presented and the discussions at the sessions of the commissions. The success of the meeting also led to certain decisions having in view the further promotion of international effort in radio research. It was determined that a campaign for funds for this purpose would be conducted, primarily with a view to permitting the general secretary's office to establish and maintain extensive bibliographical and library service, to disseminate scientific information internationally, and to serve as a clearing-house for radio data.

This organization has a distinct field of usefulness in furnishing a meeting ground for the numerous workers on the various aspects of radio research. Meetings and committee activities have furnished a means of promoting acquaintance and a meeting place of ideas. The work should be fruitful in the years just ahead in contributions to the better understanding and use of radio and in related problems of geophysics, composition of the atmosphere, etc. Observers contemplating participation in the work should, wherever possible, plan to continue for at least a year thus covering seasonal variations. It has been demonstrated that radio is unique among the fields of scientific work in having special adaptability to a large scale international research program. The phenomena that must be studied are worldwide in extent, and yet are in a measure subject to control by the experimenters.

THE TUNED-GRID, TUNED-PLATE, SELF-OSCILLATING VACUUM-TUBE CIRCUIT*

By

J. Warren Wright

(U. S. Naval Research Laboratory, Bellevue, Anacostia, D. C.)

Summary—*Certain discrepancies in published reports on this circuit are pointed out and it is shown that the frequency of oscillation is given by the real terms and the necessary condition for these oscillations to start by the imaginary terms of the general equation.*

Equations are set up for the frequency and for the necessary conditions for the generation of oscillations of a frequency determined by the tuning of the grid circuit. The feedback is assumed to be through the grid-plate capacity of the vacuum tube.

THIS is not a new circuit, and it has been discussed[1] many times in the past. Recently, however, a published report gave results which were not in accord with previous theoretical or experimental work. It seemed advisable, therefore, to investigate these discrepancies and the following article is concerned with presenting the result of this investigation in a simplified manner.

The above-mentioned report[2] stated that if the frequency of oscillation is determined by the tuning of the grid circuit, this frequency is greater than the natural frequency of the load in the plate circuit. This would mean that the reactance of the plate circuit load would be capacitative, which is contrary to the investigation of Miller,[3] who showed that the load in the plate circuit must be inductive if resistance neutralization is to take place in the grid circuit. Miller also pointed out that oscillations would occur if the resistance of the grid circuit is completely neutralized.

An examination of that report[2] showed that the effect of the feed-back current upon the total plate current was neglected. (See Fig. 1.) This is not believed to be justified; because the feed-back in such a circuit is through the grid-plate capacity of the tube,

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* Original Manuscript Received by the Institute, April 4, 1928. Published by permission of Navy Department.

[1] Oscillating audion circuits—L. A. Hazeltine, Proc. I. R. E., **6**, p. 63; April, 1918.

[2] The tuned-grid tuned-plate circuit—J. B. Dow, Proc. I. R. E., **15**, p. 397; May, 1927.

[3] Dr. John Miller, Bureau of Standards Scientific Paper No. 351.

and this current serves as the connecting link between the grid and the plate circuits.

It was also noted[2] that the frequency of oscillation was obtained from the imaginary terms and the condition for oscillation from the real terms. This is in error, for the frequency is given by the real terms and the condition for the starting of oscillations by the imaginaries.

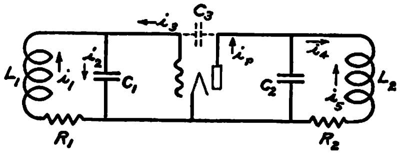

Fig. 1

This will be seen if one refers to Fig. 2 and assumes that free oscillations exist in the circuit. Then if i is the instantaneous current

$$Ri+j\omega Li+\frac{i}{j\omega C}=0 \tag{1}$$

which can be written as

$$\omega^2LC-j\omega RC-1=0 \tag{2}$$

Equating the real terms to zero

$$\omega^2=\frac{1}{LC} \tag{3}$$

It is obvious that (3) gives the frequency.

Equating the imaginary term to zero and substituting for ω

$$R\sqrt{\frac{C}{L}}=0 \tag{4}$$

for the condition for sustained oscillations.

If one remembers that the logarithmic decrement is sometimes written[4] as $\pi R\sqrt{C/L}$ it is at once obvious that $R\sqrt{C/L}$ is a damping term. In case the damping term does not equal zero[5] the oscillations will die out as shown in Fig. 3a; but if the damping term is zero the oscillations will be undamped, as shown in Fig.

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[4] Bureau of Standards Circular No. 74.

[5] (Author's Note: It is assumed that the damping term cannot be positive.)

3b. In order for (4) to hold the losses in the circuit must be supplied, and to do this in the case of the grid circuit of a vacuum-tube oscillator the load in the plate circuit must be inductive.

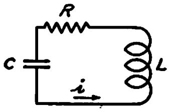

Fig. 2

In order to set up the necessary equations for the tuned-grid tuned-plate circuit refer to Fig. 1 and assume, for simplicity, that the grid does not draw current, that the tube has a linear characteristic, and consider only small oscillations. In addition, assume that the resistance of the coils L_1 and L_2 is small compared to the reactance. Then the conductance of the L_1R_1 branch is $R_1/\omega^2L_1^2$ and that of the L_2R_2 branch is $R_2/\omega^2L_2^2$. Call these conductances g_1 and g_2, respectively. The condensers C_1, C_2 and C_3 are assumed to be perfect.

Now express the various currents of Fig. 1 in terms of the product of a voltage and an admittance. Thus in the grid circuit

$$i_1 = e_g\left(g_1 - \frac{j}{\omega L_1}\right) \tag{5}$$

$$i_2 = e_g j\omega C_1 \tag{6}$$

$$i_3 = (e_p - e_g)j\omega C_3 \tag{7}$$

$$i_3 = i_1 + i_2 \tag{8}$$

Substitute (5), (6), and (7) in (8), collect terms and solve for e_p

$$e_p = \frac{e_g(g_1 + jb_1)}{j\omega C_3} \tag{9}$$

where $b_1 = [\omega(C_1 + C_3) - 1/\omega L_1]$

Similarly in the plate circuit

$$i_p = i_3 + i_4 + i_5 \tag{10}$$

$$i_4 = e_p\left(g_2 - \frac{j}{\omega L_2}\right) \tag{11}$$

$$i_5 = i_p j\omega C_2 \tag{12}$$

$$i_p = -e_g G \tag{13}$$

where G^6 is the mutual conductance of the plate circuit.

Substituting (7), (11), (12) and (13) in (10) and collecting terms

$$-e_gG=l_p(g_2+jb_2)-e_gj\omega C_3 \tag{14}$$

where $b_2=[\omega(C_2+C_3)-1/\omega L_2]$

Next substitute the value of e_p as given by (9) in (14) and arranging terms

$$Gj\omega C_3+g_1g_2+jb_1g_2+jb_2g_1-b_1b_2+\omega^2C_3^2=0 \tag{15}$$

The real terms of (15), if g_1g_2 and $\omega^2C_3^2$ (which are very small) are neglected, become

$$[(C_1+C_3)(C_2+C_3)L_1L_2]\omega^4 - [L_1(C_1+C_3)+L_2(C_2+C_3)]\omega^2+1=0 \tag{16}$$

This is a quadratic equation in ω^2 and if the theorem relating to the product and to the sum of the roots is recalled, the following can be written from inspection

$$\omega_1^2\omega_2^2=\frac{1}{L_1L_2(C_1+C_3)(C_2+C_3)} \text{ and } \omega_1^2+\omega_2^2=\frac{1}{L_1(C_1+C_3)}+\frac{1}{L_2(C_2+C_3)} \tag{17}$$

From (17) it can be seen that

$$\omega_1^2=\frac{1}{L_1(C_1+C_3)} \text{ and } \omega_2^2=\frac{1}{L_2(C_2+C_3)} \qquad (18)^7$$

Here L_1, L_2 and C_1, C_2 are the grid and plate inductances and capacities respectively. The grid-filament capacity is supposed to be a part of C_1 and the plate-filament capacity of the tube a part of C_2. C_3 is the grid-plate capacity of the tube.[8]

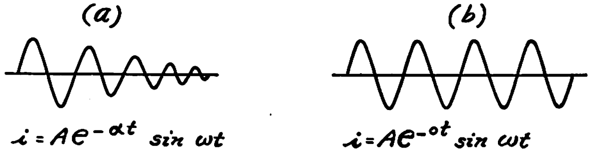

Fig. 3

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[6] Hazeltine, loc. cit., and Van der Bijl, "Thermionic Vacuum Tube," McGraw–Hill.

[7] (Author's Note: These equations are correct to a first approximation. More extended and rigorous methods give results which differ slightly but do not invalidate these conclusions.)

[8] (Author's Note: This is an approximation which is very close if C_1 and C_2 are large compared to the grid-plate capacity of the tube. This is usually true in practice.)

The condition for oscillations to occur is obtained from the imaginary terms of equation (15) or

$$G\omega C_3+b_1g_2+b_2g_1=0 \tag{19}$$

Now the condition for oscillation, when the frequency is determined by the tuning of the grid circuit, is desired, and for this case

$$\omega^2=\omega_1{}^2=\frac{1}{L_1(C_1+C_3)}$$

Then (19) becomes, on solving for G

$$G=\frac{R_1(C_1+C_3)[L_1(C_1+C_3)-L_2(C_2+C_3)]}{L_1L_2C_3}\cdot \qquad (20)^7$$

This means that the adjustment must be made so that the value of G as given by (20) is not greater than that obtainable from the vacuum tube. G will have a minimum value when

$$L_1(C_1+C_3)=L_2(C_2+C_3)$$

and as G cannot become negative

$$L_1(C_1+C_3)\geqq L_2(C_2+C_3) \tag{21}$$

but as C_3 is very small (see equation (20)) $L_1(C_1+C_3)$ cannot be much greater than $L_2(C_2+C_3)$ if G is to be of such value that oscillations are possible.

It is evident that if this circuit is to generate oscillations, of a frequency determined by the tuning of the grid circuit, the load in the plate circuit must have a natural frequency slightly greater than the frequency of the generated oscillations. The load in the plate circuit will be inductive under these conditions.

RADIATION AND INDUCTION*

By

R. R. Ramsey and Robert Dreisback

(Department of Physics, Indiana University, Bloomington, Indiana)

Summary—*In this paper the formulas for induction fields are derived in two elementary ways. From the definition of potential and by means of analogy the form of the vector potential, $A = Ih/x$ is obtained.*

From this the formulas for radiation fields are given following Dellinger's derivation. The second part of the paper gives experimental data verifying the formulas for radiation and induction when the distance is short.

THE formula for radiation from an antenna aerial and a coil has been worked out by Dellinger[1] using a vector potential. The mathematical conception of a vector potential seems to be a stumbling block. Many texts do not attempt an explanation of the radiation formulas. One text states that the development requires more mathematics than that possessed by the average engineer. Others make statements which lead to erroneous conclusions.

This paper is not intended to be a rigorous mathematical development of the theory or an improvement on Dellinger's admirable paper. It is hoped, by means of analogies using the ordinary definitions of field and potential and their application to circuits carrying direct current and without doing violence to mathematical principles, to clear up some of the difficulties which are encountered by the man who has not the time or training to look more deeply into the subject.

Radiation theory has been experimentally verified with a considerable degree of accuracy when the distances involved have been hundreds of miles. This, of course, involves an attenuation factor the uncertainty of which explains any discrepancy. There seems to have been no attempt made to verify the formula when the distance is small—one wavelength or less. In the experimental part of the paper the junior author has verified the laws of radiation and induction close to a loop aerial.

It has been customary in giving an elementary picture of radiation to picture a vertical wire with a large ball at the top for capacity. When the ball is charged positively the electric field is pictured by means of lines extending from the ball to

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* Original Manuscript Received by the Institute, November 30, 1927.

[1] Scientific Paper of the Bureau of Standards No. 354.

the earth, very much like streams of water from a spray nozzle. The magnetic field is represented by concentric circles about the wire. This elementary picture serves very well until we remember that the electric and magnetic fields as shown by Maxwell's equations for electromagnetic waves are in time phase while the fields in the picture are out of phase. The electric field extending from the ball is a maximum when the charge on the ball is a maximum, which is at a time when the current in the wire is zero, and our ordinary conception of magnetic field tells us that it is zero when the current is zero.

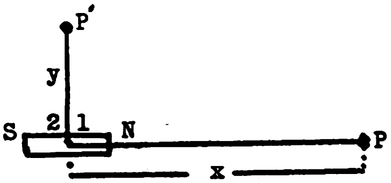

Fig. 1

This will be shown to be due to the fact that magnetic fields as we usually think of them are what have been called by Dellinger induction fields and not radiation fields.

To understand magnetic fields and potentials it will be necessary to refer to some elementary theory which is worked out in detail in such books as Starling's "Electricity and Magnetism." First we shall define magnetic potential at a point as the work[2] done in carrying a unit north magnetic pole from infinity to the point in question. The value of the potential at a point, is $V=m/d$ where m is the pole strength and d is the distance of the point P from the pole, m. In the derivation it will be seen that we first get the difference of potential between two points and then assume that the farthest point is at infinity. In fact we do what we always do when we measure potential. We find the difference of potential between two points.

A second definition of potential is a mathematical one which we shall use as the occasion arises. This definition is: Potential is a quantity whose negative space rate of variation in any direction is the strength of the field in that direction.

As an illustration we shall find the potential at a point, P, due to a short magnet (Fig. 1). Since $V=m/d$ we have the potential due to the north end as $m/(x-l)$ where $2l$ is the length of the magnet and x is the distance of P from the center of the magnet. The potential due to the south end is $m/(x+l)$. Then

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[2] (Author's Note: Strictly speaking the potential V is numerically equal to the work done in carrying unit north pole.)

$$V=m(1/(x-l)-1/(x+l))=\frac{m(x+l)-(x-l)}{(x^2-l^2)}$$

$V=2ml/(x^2-l^2)$ which when the magnet is short enough to neglect l in comparison with x, becomes $V=M/x^2$, where $2ml=M$, the magnetic moment of the magnet. We can get the field, H, by taking the negative space derivative of the potential. The field $H=-dV/dx=2M/x^3$. This follows from the second definition of potential.

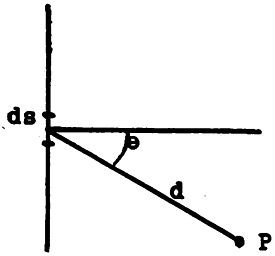

Fig. 2

If we apply the same general process as that used in deriving the potential due to a magnet, Fig. 1, the field at the point P can be found as follows: Since the field H is defined as being numerically equal to the force on a unit pole if the unit pole were at the point P, and since force is equal to the product of pole strength divided by the square of the distance,

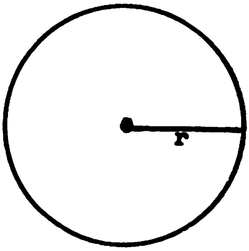

Fig. 3

$$H=m/(x-l)^2-m/(x+l)^2=\frac{m((x+l)^2-(x-l)^2)}{(x^2-l^2)^2}$$

$$=\frac{4mlx}{(x^2-l^2)^2}=\frac{2Mx}{(x^2-l^2)^2}, \text{ where } M=2ml,$$

which when the magnet is short becomes $H=2M/x^3$, which is the same as that derived above.

In the same manner for a point P', Fig. 1, placed broadside on the magnet at a distance y, the potential is $V=0$, and the field is $H=M/y^3$. Although the potential is zero the derivative of V in a direction parallel to the axis of the magnet is M/y^3 as shown below.

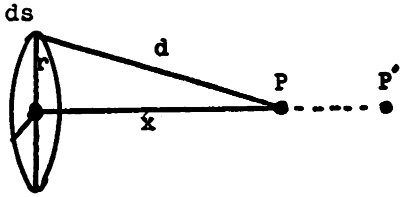

Fig. 4

The general relation of magnetic field, H, to a current in a wire may be given by the equation, $H=Ids \cos \theta/d^2$, where I is the current in an elemental length ds and θ is the angle between d and the normal to ds, Fig. 2. If this applies to a coil of wire the field at the center of the coil is $H=I2\pi r/r^2=2\pi I/r$, (Fig. 3.). If the point P is on the normal to the plane of the coil at a distance x from the coil (Fig. 4) we have

$$H=\frac{2\pi rI}{x^2+r^2}\,\frac{r}{\sqrt{x^2+r^2}}$$

which becomes, when x is great compared to the radius r,

$$H=2\pi r^2 I/x^3=2IA/x^3,$$

where A is the area of the coil. If we assume that IA is the magnetic moment of the coil we have $H=2M/x^3$, which corresponds to the value derived for a short magnet. And

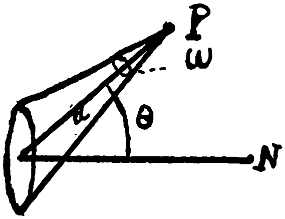

Fig. 5

$$V=M/x^2=IA/x^2=I\omega,$$

where $\omega = A/x^2$ is the solid angle at P subtended by the area of the coil. When the point P is not on the perpendicular to the coil, Fig. 5, the solid angle is $A \cos \theta/d^2$, where θ is the angle between d and the normal.

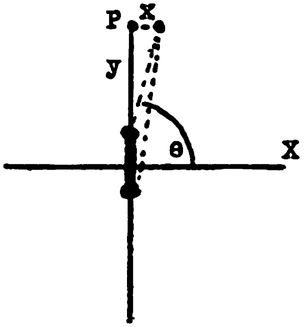

Fig. 6

In the neighborhood of the point P, Fig. 6, the potential is

$$V = IA \cos \theta/y^2 \text{ or } V = (IA/y^2)(x/y) \text{ and } H = -dV/dx = IA/y^3.$$

In the cases cited, H is the value of the field in the x direction, the direction along which the space derivative is taken which is perpendicular to the face of the coil. The potential V is scalar, that is, independent of direction and the negative derivative with respect to x gives the field along x.

If we apply the equation $H = Ids \cos \theta/x^2$ to a vertical antenna, Fig. 7, and consider the point P to be so situated that for practical purposes $\cos \theta = 1$, we have $H = Ih/x^2$, where h is the effective height of the aerial.

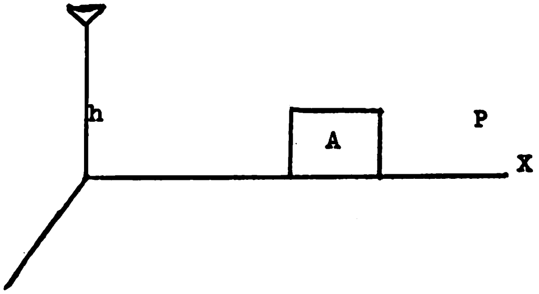

Fig. 7

Since $H = Ih/x^2$ and the potential is such that $-dV/dx = H$, then the potential must be $V = Ih/x$ since the negative derivative of this value gives the field H. In this case we know that the field H is represented by concentric circles with the aerial, h, as a center and that H at the point P is perpendicular to the direction of x. The expression Ih/x has the same form as Dellinger's vector potential, $(i)h/x$.

So far we have assumed that the current, I, is direct current or if alternating current the virtual field due to the alternating current, I, is numerically the same as that produced by a direct current of the same value. Since the current in an aerial is alternating current, I must be replaced by the expression $I=I_0 \sin \omega t$. If we use this value of I in the expression for V and take the space derivative and use the "root mean square" we get the same value for H as before. This is the field which is called induction. Thus induction is the field we usually think of when we speak of self-induction, of mutual induction, of transformers and induced currents. Due to induction, energy is stored up in the field when the current is increasing and the energy is again absorbed by the circuit when the current is decreasing. In a pure inductive circuit with alternating current no energy is dissipated. All the energy stays at home. None is radiated into space. The current is "wattless."

In the case of the aerial we know that the field at any particular instant is different at different positions. The field at a particular time at a particular point is due to the current which was in the aerial a fraction of a second before. The current may have reversed several times in the meantime. To take this into account we must change our equation for the current to $I=I_0 \sin \omega(t-t')$ where t' is the time it has taken for the wave to travel from the aerial to the point in question. Since $t'=x/v$ where v is the velocity of light our value of the vector potential is

$$V=\frac{hI_0 \sin \omega(t-x/v)}{x}.$$

Then

$$H=-\frac{dv}{dx}=\frac{d}{dx}\;\frac{(hI_0 \sin \omega(t-x/v))}{x}$$

$$H=\frac{hI_0}{x^2}\sin\,(\omega t-x/v)+\frac{hI_0\omega}{xv}\cos \omega(t-x/v).$$

Thus we see that the field, H, consists of two parts. The first is the field we get by considering the current to be constant, or if alternating current, by considering the field to be independent of the sine of the angle. This virtual field is numerically the same as the field due to a direct current.

The second part is that in which we consider the angle to depend on the distance x. The two parts are out of phase by 90 degrees. We remember we had trouble with the ordinary field in our elementary picture because it was out of phase with the electric field. This second part is in phase with the electric field.

The first part is induction. The second part is radiation. The first part, the induction, diminishes as the square of the distance while the second, the radiation, diminishes as the distance.

We can write the virtual values of the magnetic field by considering the sine and cosine to be unity, and writing I for the virtual current, then

$$\text{Induction, } H = hI/x^2$$
$$\text{Radiation, } H = hI\omega/vx.$$

If I is measured in amperes, $I/10$ will give the value of I to make the field in lines per square centimeters. Since

$$\frac{\omega}{v} = \frac{2\pi}{\lambda}$$

$$\text{Radiation, } H = \frac{hI2\pi}{10\lambda x}\cdot$$

Equating the two values and solving for x we find that the two components of H are numerically equal when $x = \lambda/2\pi$. At a distance equal to 1/6.28 of a wavelength the two values are numerically equal. Since they are in time quadrature the measured value will be 1.414 times the calculated value of one. Closer to the aerial the value of H is nearly all induction and

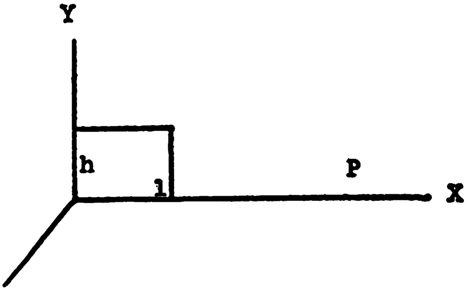

Fig. 8

diminishes as the square of the distance. Beyond this point the field is mostly all radiation and varies inversely as the distance.

If instead of an antenna aerial we have a coil aerial the induction can be calculated as is done in the first part of the paper. Induction is the ordinary field due to direct current. It is found to diminish as the cube of the distance from a coil. It is $2IA/d^3$ perpendicular to the plane of the coil and IA/d^3 in the plane of the coil.

For radiation we follow Dellinger, considering a square coil in the XY plane of height h, and length l, Fig. 8. The horizontal parts will not contribute to field at a point P in the horizontal plane. Then the radiation at P consists of two components, one each from the two vertical wires. These two will be equal but slightly out of phase because the distance of one is greater than the other by l centimeters. The resultant field at P is the vector difference of the two equal vectors which differ in direction by a small angle, θ. $\theta/2\pi = l/\lambda$ or $\theta = 2\pi l/\lambda$. In the diagram, Fig. 9, $oa = 2H_1 \sin \theta/2$. Since θ is small, $\sin \theta/2 = \theta/2$, then

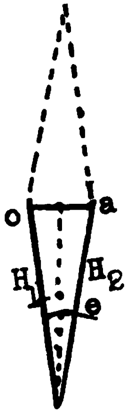

Fig. 9

$$oa = H = 2\pi\left(\frac{hI}{\lambda x}\right)\frac{2\pi l}{\lambda} = 4\pi^2 hlI/\lambda^2 x.$$

Thus the radiation from a coil varies inversely as the distance while the induction varies inversely as the cube of the distance.

The radiation from a coil varies inversely as the square of the wavelength while from an antenna inversely as the wavelength.

The induction from a coil is $1A/d^3$ and the radiation from the same coil is $4\pi^2 IA/\lambda^2 d$. Equating the two values we get $1/d^2 = 4\pi^2/\lambda^2$ or $d = \lambda/2\pi$. Thus the two components are equal at a distance $\lambda/6.28$, the same being true for an antenna aerial. Close to the coil or antenna aerial the field is primarily induction.

For practical purposes when the distance is less than 1/20 of a wavelength the radiation can be neglected, and when the distance is greater than 1/2 wavelength the induction can be neglected.

The energy represented by induction does not leave the aerial. It is stored in the medium during the first fourth of a cycle and then returns to the aerial during the second fourth of the cycle in the same manner as the field of an ordinary transformer or choke coil. The induction is the field which stays at home. The energy of the field of the radiation does not return to the aerial but passes out to infinity unless it is absorbed by intervening objects. The energy is radiated into space.

Of course it is possible to absorb a part of the energy of induction if the absorber is in the field of the induction, that is, near the aerial. This is the same as in a transformer part of the energy may be absorbed by the secondary coil in which case it cannot return to the primary.

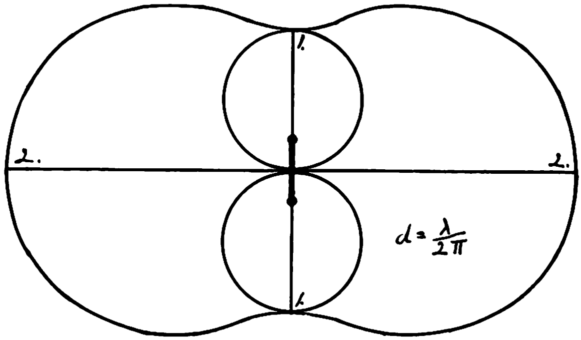

Fig. 10—Showing the Distribution of Radiation and Induction when the distance $d = \lambda/2\pi$. In the plane of the coil the radiation and induction are equal. Perpendicular to the coil the radiation is zero while the induction is twice the value in the coil's plane.

It will be noted that radiation from a given aerial depends upon the frequency or wavelength. Induction is independent of frequency. The virtual value of induction field for 60 cycle, 300 meters or 41 meters is numerically the same as that produced by direct current.

The induced emf in a vertical receiving aerial is equal to the number of lines of force cut per second. This for a magnetic field of H moving with a velocity of v (velocity of light) centimeters per second past a receiving antenna of height h_r is $e = h_r H v$ absolute units of electromotive force. Since in an electromagnetic wave there is always an electric field, E, perpendicular to the magnetic field, H, the emf in the antenna is $e = E h_r$, E being expressed in absolute units per centimeter. Since the induced electromotive force is the same in both cases $e = E h_r = h_r H v$, and $E = Hv = 3 \times 10^{10} H$.

It turns out that if E is expressed in microvolts per meter the numerical value is the same as if expressed in absolute units per centimeter.

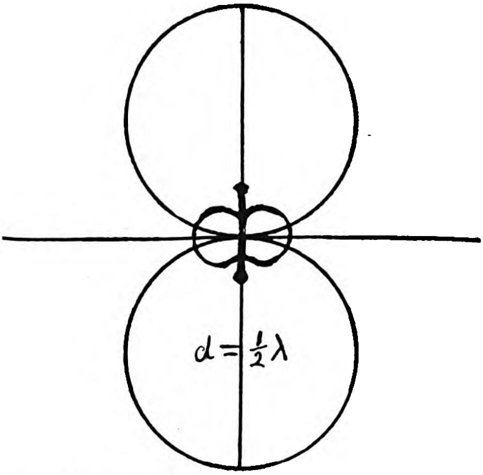

Fig. 11—Showing the Distribution of Radiation and Induction when the distance $d = \frac{1}{2}\lambda$. The induction can be neglected.

If a coil is used as receiver the induced emf is calculated from the rate of change of the field through the coil, Fig. 7.

Thus $E = H A_r 2\pi n / 10^8$ volts.

The effective height of a coil is the height, h, of a receiving antenna in which the theoretical induced emf is the same as that in the coil for the particular radiation. This height changes with frequency. The higher the frequency the greater the effective height of a particular coil.

From the fundamental equations as given above, Dellinger's four equations for received current in coils and antennas can be derived. Since they are well-known they do not need to be repeated here.

To sum up the above, the potential and field due to a short magnet and to a coil have been derived, and from a rather loose analogy the vector potential has been derived in the case of an antenna. From this vector potential the radiation has been derived.

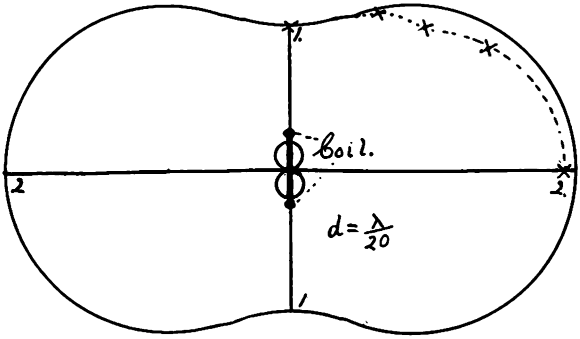

Fig. 12—Showing the Distribution of Induction about a coil according to the equation $H = IA/d^3\sqrt{1+3\cos^2\theta}$ where θ is measured from the normal to the coil. The dotted line shows the measured values. The two small circles at the center show the relative values of the radiation when $d = \lambda/20$.

Experimental Results. In the above formula it will be noted that the shorter the wavelength or greater the frequency the greater the radiation field from a coil or antenna. And since the received current in a coil increases with frequency, it is advantageous to use short waves. Since the height of an antenna is an uncertain quantity, a coil was used. The frequency selected compared to a wavelength of 16 meters. Short waves are an advantage since the radiation and induction are numerically equal at $\lambda/2\pi$ and at a distance greater than 1/2 wavelength the induction field can be neglected.

Fig. 10 shows the relative distribution of radiation and induction when the distance is $\lambda/2\pi$. The circles show the radiation, and the "squeezed" elliptical figure shows the induction. The values are numerically equal in the plane of the coil.

Fig. 11 shows the relative values of radiation and induction when the distance is $\lambda/2$. In this case the small figure-eight figure shows the induction which can be neglected.

Fig. 12 shows the distribution when the distance is $\lambda/20$. In this case the radiation can be neglected.

The radiating coil used in radiation determinations was a square coil of one turn whose area was 3140 square centimeters connected in a Colpitts circuit and made to oscillate at 16 meters. Fig. 13 shows the diagram of hook-up. The receiving coil was a single-turn coil, area 1913 square centimeters. A thermojunction was connected in the circuit and a micro-ammeter was used to register the current. The maximum received current was about 80 milliamperes and the minimum was about 5 milliamperes. The resistance of the receiving coil was 3.2 ohms.

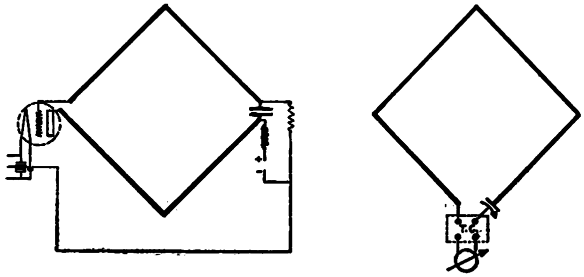

Fig. 13—Diagram of Oscillator Circuit and Receiving Circuit Used in Radiation Experiments. The wavelength was 16 meters.

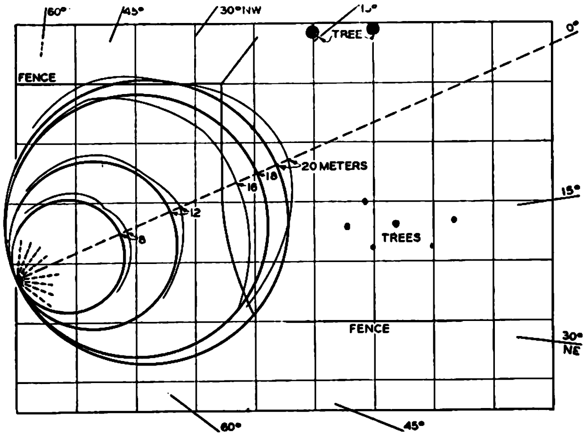

Fig. 14—Showing the General Lay of the Ground and Obstructions Such as Trees and Fence.

The heavy circles show the theoretical distribution of radiation when *d* is 8, 12, 18, and 20 meters. The lighter lines show the measured distribution.

Two fifty-watt tubes were used in the oscillating circuit. The filament supply was a 12-volt 60-cycle transformer regulated by a rheostat in the 110-volt side. The plate supply was a

1500-volt transformer with a rheostat in the 110-volt supply. The tubes were the only rectifier used. This variable supply caused the wave to be rather broad. The decrement factor was determined by connecting the tubes to a high potential d.c. generator and adjusting the circuit until the normal oscillating current was the normal value, about 5 amperes, and comparing the received current with that received when the same value, 5 amperes, was generated with the a.c. supply, the position of transmitter and receiver being the same in both cases. The ratio of the two received currents was found to be 1.2. This factor was used to correct the received current. The transmitter and transformers were mounted on a table. The only supply line was a long lamp cord through which the 110-volt alternating current was conducted. Radio-frequency chokes were placed in the supply. The simplicity and safety of this was thought to outweigh any advantage that a high-potential, d.c. supply might have.

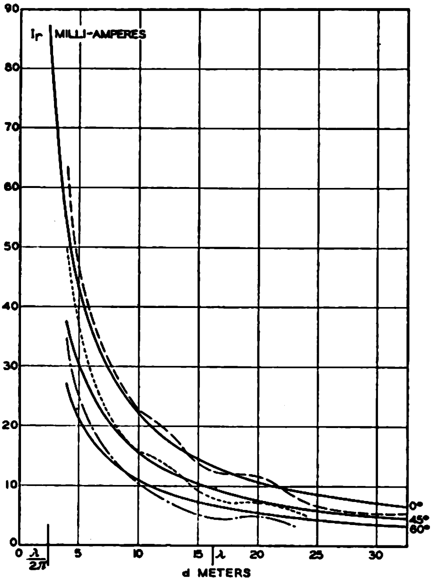

Fig. 15—Showing the Relative Values of Radiation in Terms of Received Current, with Distance for Various Angles. The angle is measured from the plane of the coil. The heavy line shows the theoretical value of the current and the broken lines the observed current.

The location was a comparatively open space in front of Science Hall, Indiana University. This was not ideal, since at points removed some distance from the oscillator the reflection from trees and other objects could be easily noticed. A more open space is desirable but the question of a convenient power supply led to the choosing of this location. A small island or rock in a lake would be ideal, a boat being used to carry the receiver.

Fig. 14 gives the general contour of the plot. The northwest quadrant was relatively free from reflections and the results check with theory for distances about 10 to 20 meters from the oscillator. Closer than 10 meters the induction was an appreciable quantity and at greater distances than 20 meters the reflections from objects were appreciable. Fig. 14 gives the theoretical distribution of the radiation field and the measured distribution. The theoretical field is shown by heavy lined circles and the measured values by the lighter lines. A guard fence of a single iron wire and several trees are also shown in Fig. 14. Figs. 15 and 16 also show the distribution in another way. Fig. 15 shows by the heavy curve the theoretical distribution which varies inversely as the distance for angles of 0 deg., 45 deg., and 60 deg. from the plane of the transmitting coil. The broken lines show the measured values. Fig. 16 shows the same thing for 15 deg. and 30 deg. Values in this case are given for measurements in the northeast

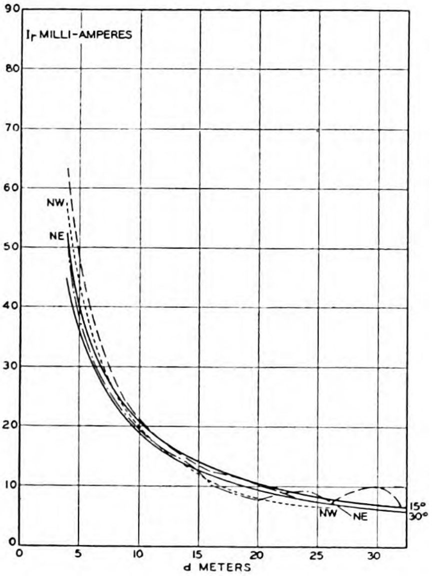

Fig. 16

and northwest quadrants. The periodic variations in the measured results are without doubt due to reflections.

For induction fields the measurements must be made so close to the aerial that radiation does not make an appreciable reading. This must be close to the aerial but not too close, since for short distances there is a question as to the point of the coil from which measurements of the distance are to be made. The distance must be great compared with the diameter of the coils. On this account short waves are not advantageous.

An oscillator of seven turns connected in the Hartley circuit and adjusted to oscillate at 500 meters was used. Fig. 12 gives the theoretical variation of the induction field with the angle. The irregular flattened elliptical figure gives the induction; the two circles give the radiation distribution. The broken line gives the measured distribution.

The measured values do not agree in every detail, but all the discrepancies can be explained in terms of reflections or reradiation from trees or other objects. The distribution of radiation about a tree makes an interesting study in itself.

DEVELOPMENT OF A SYSTEM OF LINE POWER FOR RADIO*

By

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(Conner Crouse Corporation, New York City)

Summary—A short history of the development of power supply apparatus for radio receivers is given, including a brief discussion of the technical requirements, such as the most desirable suppression characteristics, the necessity for voltage regulation, etc.

The various devices which have been developed are described.

The present status of the art and the scope and limitations of the different types of apparatus are discussed.

The series wired radio set with common power supply for plate filament and grid circuit is discussed in some detail, and curves of the operating characteristics of such a set are given.

THE history of the development of radio socket power is interesting, not alone from the technical standpoint, but also as teaching a number of valuable commercial lessons. This interest from either viewpoint arises largely from the fact that while most commercial devices stand very largely on their own feet, the radio socket power has to be coordinated in a peculiar and unusual degree with the radio receiver which it serves. Furthermore, socket powers are a combination of two elements, a rectifier and a filter. The problems in the design and manufacture of these two elements are so different that their development has always been in separate hands. Because of this separation the development of the complete unit was not in the early days well-coordinated and the influence of the separation has been felt throughout the entire history of the art.

But despite the difficulties, both natural and artificial, which surrounded the development of this art, we have now reached the point where radio, operated from the electric light socket, is an established fact. Socket-power sets are now available to the public at reasonable prices, which are reliable in operation and satisfactory in results.

When radio socket powers first appeared on the market about three years ago the regenerative circuit was almost universally used with reproducers of the head-set type and because of the comparatively small number and low power of the transmitting

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stations it was necessary to reach out for distance to get full and satisfactory programs. Because of the sensitivity of the head set and because of the low value of the available signal strength, the problem of filtering was all-important. The design of filter networks of the type required for this service was not well understood. While the mathematics of band-pass filters for telephone work had been very thoroughly worked out very little was known by radio engineers in general about the design of networks for the suppression of alternating-current components in a direct-current line.

The two problems are quite different. Let us digress for a moment to discuss the differences, for the sake of clarity.

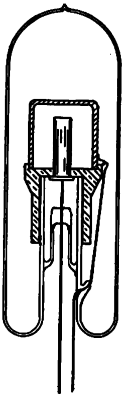

Fig. 1—"Karbo" Tube without Base.

In the first place, the telephone filter is "frequency-selective." That is, its function is to pass certain frequencies with as little loss as possible, and to suppress completely, all other frequencies. The socket-power filter, on the other hand, must very thoroughly suppress every alternating component which may be present in the output of the rectifier. These frequencies cover an enormous band. In devices operating from sixty-cycle supply, and using full-wave rectification, there will be present a large component of 120 cycles, a small 60-cycle component, and long series of higher harmonics, extending as far as the sixteenth. Of these components of course the 120-cycle is by far the largest, and it ordinarily requires the greatest suppression.

Another difference between the two types is the presence of direct current in large amounts in the socket-power filter. This means that the inductances must be of special design, and that any shunt capacities employed have a dual function, acting not only as capacities in the mesh, but as direct-current energy-storing devices.

The obvious solution was found to be the use of very high capacities, which were obtained at low cost by means of electrolytic condensers. Furthermore, since the problem of filtering is made less difficult by a decrease in the direct current, means were worked out for operating the filaments of the radio set in series.

A successful device along this line appeared on the market in the fall of 1924. The rectifiers, two in number, were of the cold cathode, gas-filled type. In Fig. 1 a line drawing of these early rectifiers is shown. In Fig. 2 is shown a diagram of the complete unit and its suppression characteristic.

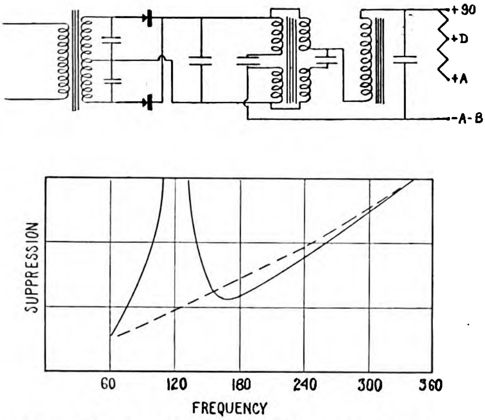

Fig. 2—Original Socket-Power and Suppression Characteristic.

In the filter of this unit appeared a principle which we have retained to the present time. As stated above, the 120-cycle component is by far the largest in amplitude, and because of its low frequency it is difficult to filter. The filter of Fig. 2 is designed to exert its maximum suppression at this frequency. For comparison, there is shown in dotted lines the suppression characteristic of the usual type of filter, using series inductance and shunt capacity. The great advantage of the former curve will be readily appreciated.

The difficulties encountered with this first unit had to do principally with the electrolytic condensers. These devices, which because of their low cost per unit of capacity appear very attractive, have the serious defect that their operation depends upon a very high degree of chemical purity of all of the substances used in their manufacture. They are particularly sensitive to chlorine and sulphur ions, and most extreme precautions must be taken in the manufacture to exclude these substances and to prevent their entrance at any time during the life of the cell.

These difficulties with the electrolytic condenser appeared early, necessitating intensive work on the theoretical and practical study of filter meshes, with the result that very shortly the electrolytic devices were replaced by paper condensers of much smaller total capacity. When this change was made the result was a complete socket-power device capable of delivering *A*, *B*, and *C* power to a set of any size, and while many improvements in details have been made since then, this early device is still commercial and could be marketed today. However, the opposition to the use of series wiring of filaments on the part of radio manufacturers has prevented the introduction of this system until two seasons ago.

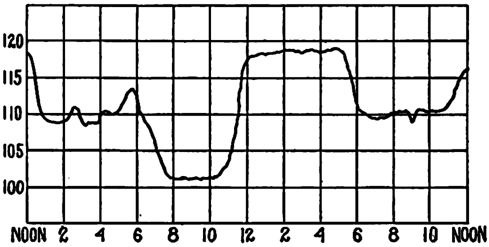

Fig. 3—Typical Line Voltage Chart.

As a result of this opposition, socket-power manufacturers had to be content with turning out a device for supplying *B* power alone, which theoretically at least could be applied to any radio set. Most of the early models of the *B*-power units, however, had very poor performance. The manufacturers had no knowledge of the demands which their devices placed on the rectifier tube, with the result that this replaceable element generally had a very short life. At the same time radio sets were in general exceedingly sensitive to inter-stage coupling, in both the radio and audio stages, and the *B*-battery unit introduced a larger inter-stage coupling than the dry battery. Despite these difficulties, however, the public eagerly absorbed everything that was put out.

While the technical knowledge and skill required to build suitable rectifiers were available at that time, it was some time before the tube manufacturers could be brought to realize the possibilities of this branch of industry. However, they have nobly made up for their lack of early effort, and now exceedingly satisfactory rectifiers are availabe for practically every class of service.

TABLE I
RECTIFIERS AVAILABLE IN 1924

Name	Description	Direct Current	D. C. Volts
Tungar	Hot Cathode Gas-Filled Half Wave	2 Amps.	40–50
Tungar	"	5 "	80
Karbo	Cold Cathode Gas-Filled Half Wave	15 "	180
———	Various Types of Electrolytic Devices		

An idea of how thoroughly this work has been done may be gained from consideration of Tables I and II. Table I gives the characteristics of the types of rectifiers available in 1924. Table II is a similar table of the various types of present-day rectifiers. It will be seen that today, whatever voltage or current the socket-power designer may require, he can find an economical rectifier commercially available for his purpose.

TABLE II
RECTIFIERS AVAILABLE IN 1928

Name	Description		Direct Current	D. C. Volts
Tungar	Hot Cathode Gas-Filled	Half Wave	2.0	40
		" "	5.0	80
		" "	0.6	100
		Full "	5.0	30
Radiotron	Hot Cathode High Vacuum	Full Wave	Up to 0.125	550
Raytheon	Cold Cathode Gas-Filled	Full Wave	0.85	300
			0.350	320
———	Crystal Type Full Wave		5.00	8–10

In the meantime the character of radio reception was changing; new and larger broadcasting stations were being built and the radio listener became more and more interested in the entertainment value of radio. The head set was replaced by the loudspeaker. This change naturally reacted on the design of the socket-power units. Because of the lower sensitivity of the loudspeaker and of the high available signal strength, the problem of residual ripple became less important, and this, together with the continually increasing knowledge of filter design, made the socket-power devices smaller and more economical. However, a new problem now arose. In the old days when the listener sat immediately in front of the set, his hands continually poised to readjust several controls, it made very little difference whether the voltages applied to the set were constant or not. The users of loudspeakers, however, desire to adjust the set and then leave it. Since the radio-set designers had found that the most satisfactory volume control was one which varied the electron emission in one or more radio stages, and since the voltage at most sockets varies through several volts, as will be seen from a typical voltage record in Fig. 3, it became absolutely essential to develop cheap and reliable voltage-control devices. Two successful types of regulator which are cheap to build, economical in operation, and reliable, are shown in Figs. 4 and 5. The device shown in Fig. 4 employs a temperature-variant resistance, while the device shown in Fig. 5 is purely magnetic.

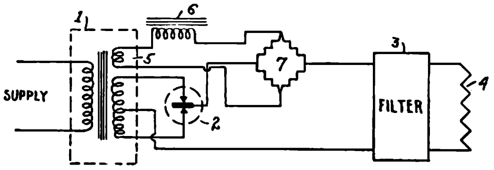

Fig. 4—Resistance Type Voltage Regulator.

In Fig. 4 the parts included in block 1 comprise a transformer; 2 is a rectifier, 3 the filter, and 4 the load. The voltage regulator comprises three elements: an additional secondary winding 5 on the transformer, an iron core inductance 6, whose core is magnetically saturated, and a temperature-variant resistance 7, arranged in the form of a bridge inserted in one of the leads connecting the rectifier to the filter.

The operation of the device may be briefly described as follows: Suppose the voltage of the supply to rise above some normal minimum value. This will cause an increase in voltage across the terminals of the secondary 5, and because of the presence of the saturated core inductance 6 in this circuit, a considerably increased alternating current will be caused to flow from the secondary 5 through the bridge resistance 7. Because of the positive temperature coefficient of this resistance its value will be much increased and will thereby absorb the additional potential across the rectifier output caused by the increased supply voltage Owing to the balanced bridge arrangement of the resistance 7, the direct current to the filter is electrically isolated from the alternating regulating current supplied by the secondary 5, so that no direct current will be wasted in the regulating circuit and no alternating components will be introduced by the regulating circuit into the d.c. circuit.

This device is, of course, not instantaneous in action, but by proper mechanical design of the resistance 7 may be made to operate in a few seconds.

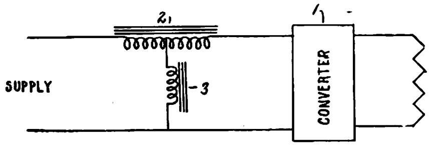

Fig. 5—Magnetic Type Voltage Regulator.

The second device, shown in Fig. 5, operates entirely in the primary side of the a.c. circuit, and the transformer, rectifier, and filter are all included in the block 1 marked "Converter." The regulator comprises a series inductive element 2, so designed that its iron core never becomes saturated throughout the normal working range of the device, and a shunt inductive element 3 whose iron core is normally saturated. The theory of operation of the device is somewhat involved, but it may be roughly stated that an increased supply voltage causes a magnified increase of current through the saturated element 3. This magnified current flowing through a portion of the inductance 2 causes an increased potential drop across that element to affect regulation, and this increase of potential drop is multiplied by the ratio of total turns to the turns included between the supply and the connection of the coil 3.

This device is instantaneous in operation and, as will have been noted, requires no temperature-variant resistance.

The great number of sets in service, together with the opposition of the manufacturers to series filament wiring, gave considerable impetus to research work directed toward the design of a filter which would supply 1½ to 2 amperes of *A* current to radio sets with parallel filament wiring. One or two abortive attempts along this line found their way to the market in 1925. These devices were inductance-resistance types, employing series inductance and shunt resistance elements. Such an arrangement has two serious disadvantages; in the first place, the amount of current which must be by-passed through the shunt elements is very large, and, secondly, its suppression characteristic is unsuited to the filtering of rectified current.

The problem was finally solved, as will be discussed more fully later, and successful filters of this type are now in operation.

We may go back at this point and trace the development of the so-called a.c. tube. Obviously, the current flowing in the filament of an audion has no function other than to heat the filament to the point where emission in satisfactory quantities takes place. It would therefore seem that since alternating current is just as satisfactory for heating purposes as direct current, it would be possible to use alternating current directly on the filaments. Alternating current had been so used as early as 1916 to heat filaments of large audions in transmitting stations. A part of the difficulty arising from the variation in potential drop across the filament (which variation is introduced into the grid circuit) was overcome by the scheme proposed, apparently independently, by White and Heising, of connecting the grid and plate returns to the center point of the filament, or the center point of an impedance connected across the filament. For transmitting tubes this worked very well, but when it was attempted to apply it to radio receivers where the signal energy was very small, it was found that even with the midpoint connection both radio and audio difficulties developed. It was at first thought that these difficulties were connected in some way with variations in filament temperature at different parts of the a.c. cycle. In the early part of 1923 the writer performed a series of experiments with tubes having wide variation of ratio of filament mass to filament radiating area, which showed conclusively that the difficulty was in no way connected with the alternate heating and cooling of the filament, except with very fine filaments such as were used in the 60-milliampere tubes. The trouble was caused entirely by the varying drop across the filament. Tubes in which the drop has been reduced to 1½ volts are now on the market and their sphere of usefulness will be discussed a little later.

With the investigation of these phenomena another method of solution presented itself. I am not sure where the suggestion originated; however, in 1919 Professor Morecroft described in his book, "Radio Communication," an equi-potential cathode tube in which alternating current might be used to heat an element which in turn radiated heat on to a cathode in no way connected to the a.c. circuit. This line of development has also been brought to a commercial point and several manufacturers have placed such tubes on the market within the past year. These tubes are much more satisfactory from the radio designer's standpoint but seem to be inherently expensive to build.

Having taken a brief view of the history of line-power device, we now turn to a discussion of the various types of apparatus available to the radio-set manufacturer and the public, and the scope of usefulness of each type.

The a.c. tube has for some reason proved very alluring to radio-set manufacturers. Its principal advantage seems to be compactness. A number of sets have appeared employing an indirect heater type of tube in the radio socket and the low-voltage cathode type in the other sockets. It is too early to say much concerning the practical performance of these sets. It seems fairly definitely established, however, that in the present state of development of these tubes it is not possible to build as good a radio or audio amplifier as with the d.c. type. Sets employing indirect heaters in all stages may be made to give high quality of reproduction, but this greatly increases both the first cost and the cost of replacement.

Another of the principal difficulties met with in the a.c. tubes has been their sensitiveness to the filament excitation voltage. Since wide variations of line voltage are encountered in the United States this has proved very troublesome. The only satisfactory solution seems to be the inclusion of a voltage regulating device as a part of the radio receiver. Since at least one cheap and satisfactory device for voltage regulation is now available, it seems quite probable that this step will be taken on a large scale in the near future.

The parallel filament device with a suitable *B* socket power has its field of usefulness in connection with high-grade sets and also as an accessory for converting sets now in the users' hands.

Turning to the first mentioned field, the radio-set manufacturer is beginning to realize that it is very costly to change the design of his product radically every season. The money invested in special tools is obviously thrown away. What is not generally appreciated is that the money invested in the time of the engineers is also thrown away. It may of course be said that engineers are a necessary part of the radio manufacturers' organization and may as well be kept occupied, but it seems probable that at the present time research work along the line of detail improvement is more profitable than the design of an entirely new receiver every year.

The large current socket power furnishes a means of converting a d.c. radio set into a socket-power set, without any change in the radio set itself; so far as the high-priced complicated units are concerned, this has proved very attractive.

Two types of units are available for this service. In Fig. 8 is shown a diagram of one electrical type, which is already in successful commercial operation.

In Fig. 6 the voltage regulator is represented by 1—the voltage regulator was described previously,—2 is the transformer, and 3 the rectifier, of the Tungar full-wave type. The filter, shown in 4, is of the inductance-resistance type. What may be termed the "Siamese System" comprises the three coils 5, 6, and 7, and the band 8, all mounted as shown on the two adjacent iron cores 9 and 10. Two resistances are employed, one, 11, connected across the coil 5, and the other, 12, connected between the junction point of the coils 5 and 6, and the negative side of the d.c. line. This device has some interesting and valuable properties. It is a "blanket filter," having a high degree of suppression at all frequencies, and, curiously, although it employs only resistance and inductance, it has a distinct "anti-resonance" point which by suitable design may be made to fall at 120 cycles, or at any other desired point.

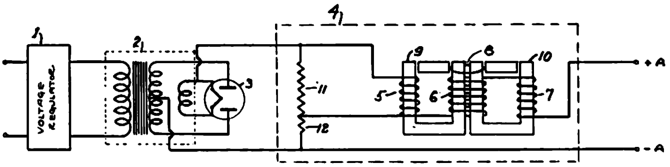

Fig. 6—High-Current Socket Power.

The total inductance of the device, under full d.c. load, is less than one henry, but the r.m.s. value of the unfiltered ripple at the output is less than one-tenth of one per cent. This high filtering efficiency with low values of inductance is very desirable. It should be emphasized that, particularly in high-current filters for this class of service, the physical size of the inductance units is not determined by the temperature to which the insulation may be worked, as in most electrical apparatus. The size is determined by the effect of the change in temperature of the coils on their resistance, which in turn causes a change in the output voltage as the device heats up. For this reason the coils must be worked at a very low temperature. This makes them very large and expensive, unless, as in this case, special circuits are employed to increase their effectiveness.

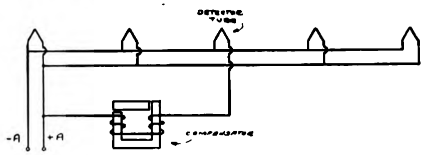

Fig. 7—Detector Filter.

Where the socket-power unit is intended for use with a particular set, its size and cost may be further reduced by the use of the simple arrangement shown in Fig. 7. It is well-known, of course, that the residual unfiltered ripple in the output of the socket power is amplified by the audio amplifier of the set, and it is therefore obvious that this ripple is more troublesome in the filament of the detector tube than anywhere else. By the simple use of the small coil shown in the figure, inserted in the filament circuit of the detector tube, the filaments of all of the other tubes of the set are made to act as the shunt resistance element of a one-section filter added to the filter of the socket-power device, for the benefit of the detector tube.

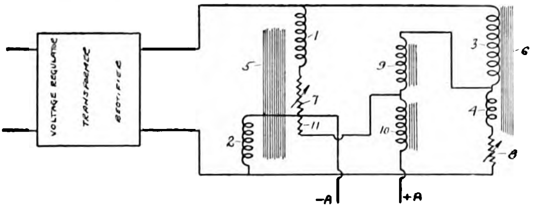

Fig. 8—Bridge-Type *A* Socket Power.

In Fig. 8 is shown a bridge type of filter for the same class of service. In this figure, the filter only is shown, the details of the voltage regulator, transformer, and rectifier being the same as in Fig. 6. The filter comprises two sections, the first being a bridge section, formed by the coils 1, 2, 3, and 4 wound on the cores 5 and 6 as shown, and the resistances 7 and 8; and a *T* section, formed by the two inductances 9 and 10 and the resistance 11. The bridge is balanced to the most troublesome frequency of the rectifier output and completely suppresses this frequency. At the same time the bridge, by the peculiar arrangement of the elements and by the use of the "reversed mutual" inductance between the coils 1 and 2 and 3 and 4, has a fairly good suppression characteristic at all higher frequencies. This has a double advantage in that it leaves very little to be done at the higher frequencies by the *T* section, and at the same time makes the bridge balance insensitive to change of frequency. A reduction in size of this unit may also be effected by the use of the arrangement shown in Fig. 7.

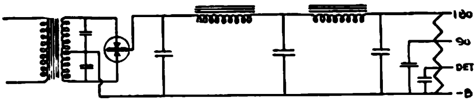

Fig. 9—Usual *B* Circuit.

Both of the units just described may be used as accessories. That is, they may be applied to sets now in service. They are dry, have nothing to wear out, with the exception of the rectifier, and can be marketed at a reasonable list price. It is not probable that those who now own high-grade sets operated from batteries will be willing lightly to dispose of them, and these units are a solution of that problem.

Within the last few months, a new type of electrolytic condenser for use in low-voltage, high-current filters has been incorporated in at least two socket-power devices now on the market, but no operating data are yet available on them.

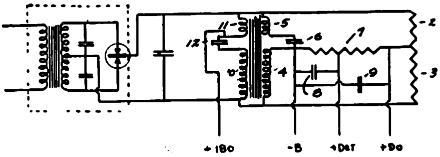

Fig. 10—Improved *B* Circuit.

The filament-supply devices just described may be used with *B* batteries, or with any form of *B* socket-power device. The most commonly used circuit is that shown in Fig. 9, and is too well-known to need description. In Fig. 10 is shown a type of *B* socket-power which presents a number of advantages over the older form. In this figure, the transformer and rectifier are shown in block 1. The filter for all of the amplified stages except the last is of the bridge type, two arms being formed by the resistances 2 and 3, one arm by the inductance winding 4, and another arm by the inductance 5 and the condenser 6. The detector voltage is reduced to the proper value by means of the resistor 7, and the usual signal by-pass condensers 8 and 9 are provided. For the last tube, the two resistive arms are replaced by the arms formed by the coils 10 and 11 and the capacity 12. For a given size of parts (and therefore, for a given cost) this filter has much better filtering efficiency and better voltage characteristics under change of load than the older type.

What appears to be a very satisfactory solution of the socket-power problem for the radio-set builder is the series wired filament set, using a single rectifier and filter for supplying *A*, *B*, and *C* power.

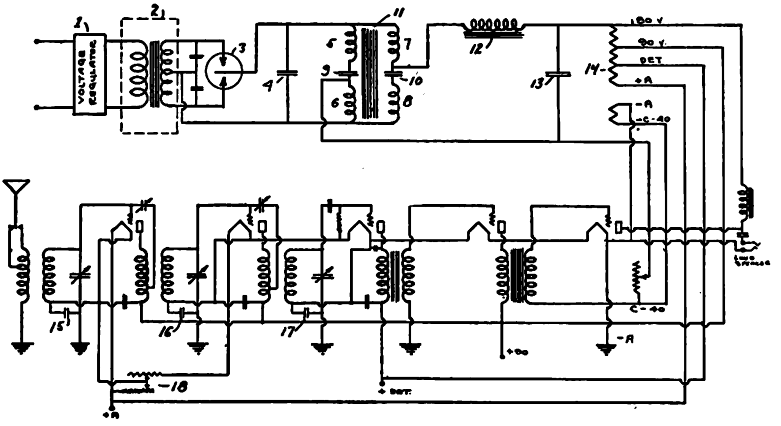

Fig. 11—Series Set—Socket Power.

In Fig. 11 is shown a complete diagram of the socket-power device and the radio set. In this figure, 1 is the voltage regulator, which may be of the type shown in Fig. 5. The parts in block 2 compose the transformer. The full-wave rectifier is shown at 3, working into the condenser 4. The filter is composed of the bridge section comprising the four coils 5, 6, 7, and 8, and the two condensers 9 and 10. This bridge is designed to be balanced at 120 cycles for a 60-cycle supply circuit. The coils of the bridge are all wound on a single iron core 11. An additional iron core inductance 12 is provided, and the entire reactive network is terminated in the capacity 13.

For sets using the 171-A tube in the last stage, the socket power is designed to deliver approximately 300 milliamperes of filtered direct current at the terminals of the condenser 13, at a potential of 220 volts, the various voltages being tapped off from the potentiometer 14, as shown.

The very small number of changes that have been made in the radio set to accommodate the series wiring of the filaments should be especially noted. In order that it may not be necessary to insulate one variable condenser from another, paper condensers have been used at 15, 16, and 17. The volume control 18 has been changed to the double type shown. These are the only changes that are necessary, and they add but little to the cost, in no way making any change in the operating characteristics of the set.

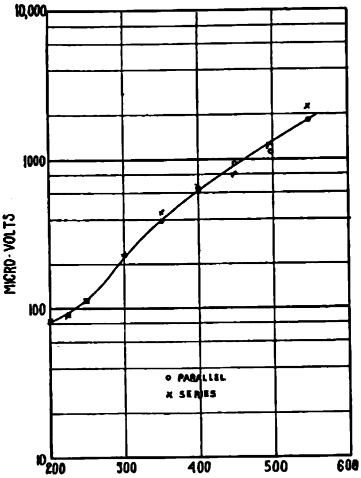

Fig. 12—Sensitivity Comparison, Parallel and Series.

A very careful check of this latter point was recently made. A high-grade receiver, built by a well-known manufacturer, was selected, and its characteristics with battery supply determined by the Radio Frequency Laboratories of Boonton. They tested it for sensitivity, selectivity, and fidelity, the results of these tests being shown respectively on Figs. 12, 13, and 14 in dotted lines. The set was then re-wired in series, according to the diagram of Fig. 11, and retested with the socket power. The results of these tests are shown in the full lines on the same figures. It will be seen that no appreciable difference in any of these characteristics is apparent.

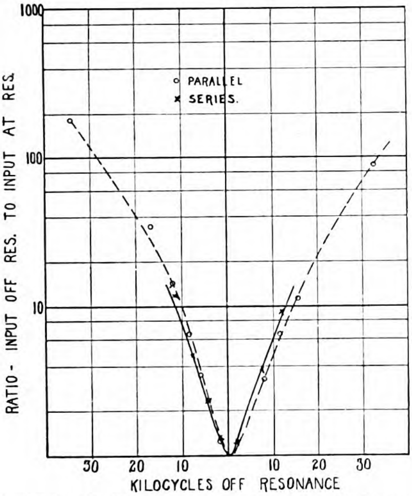

Fig. 13—Selectivity Comparison, Parallel and Series.

These tests show conclusively that the use of series wiring of the filaments with d.c. tubes, together with a socket-power device, imposes no new limitations on the radio-set designer. He does not have to use an inferior method of volume control; he can use the old reliable filament control of the r.f. stages. He does not have to sacrifice audio quality. He can use all his past experience, and all his standard parts.

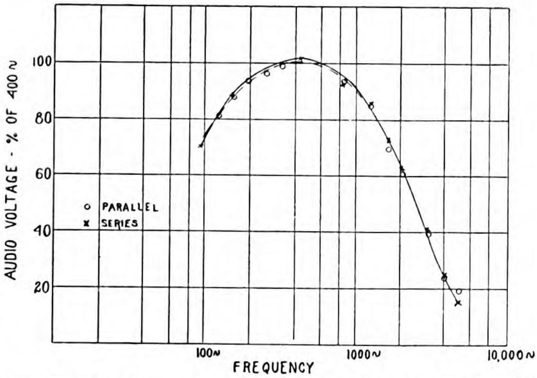

Fig. 14—Comparison of Fidelity, Parallel and Series Wiring.

In cost of construction the system is more economical than anything which has so far been proposed. Practically no expense is added to the receiver, and the socket-power device costs but little more than the *B* socket power which is a necessary part of any system. It will be noted that the socket power shown in the figure employs the same number of parts required for a *B* socket power alone. The transformer and coils are slightly larger than in a *B* device, whereas the requisite amount of capacity is usually somewhat less.

Another item of cost must be considered when comparing this system to, say, the a.c. tube system. Means have been worked out whereby the wiring of the set may be made universal. That is, the same set may be used interchangeably as a parallel-wired, battery-operated set, or as a series-wired, socket-power set. The simplification thus gained in all of the problems of manufacture and distribution hardly needs pointing out.

The development of this art to its present state has been the work of four years. We now find the major technical problems solved, and while the commercial situation is in general anything but satisfactory, we have every reason to believe that this difficulty also will be overcome in the near future.

BOOK REVIEW

History of Radio Telegraphy and Telephony, by G. G. Blake. Published by Chapman and Hall, Ltd., London. 440 pages, 270 illustrations. Price 25 shillings, net.

This book is not strictly a "history," but might more logically be called an "encyclopedia," containing as it does references to hundreds of inventions and developments pertaining to radio and associated sciences, which have been disclosed during the last half century. These items are roughly listed under sixteen chapter headings with the name of the inventor in each case displayed in bold face type.

In general the fundamental idea only of each of the items is set down in a brief summary, in non-technical language, without mathematics or other scientific proof. For this reason it makes an excellent reference book for those who have recently taken up the study of radio and are unacquainted with the many now little known steps of progress in radio technic. The old timers as well will find in the book many interesting processes described which have been all but forgotten now.

The reviewer (although acquainted with nearly all of the histories, etc., since the time he was awed with his first radio book—the first edition, then current, of Fleming's "Principles") found a great number of ideas and processes new to him. The inclusion of so many of these unique items contributes the greatest value to the book, since many of them are based on obscure physical phenomena and may again prove to be adaptable to other fields, such as television, broadcasting, etc.

It should not be gathered from the above that no present day appliances are included, for they are fairly well represented. The layman might experience difficulty, however, in distinguishing the items of present day equipment from those which are to be found in the museums. This classification, however, is so elastic, varying not only with time but with the country in which used, that this is not a defect.

One of the best features is the inclusion of a reference to the very extensive bibliography, with each item described. This bibliography contains over eleven hundred items so that the reader can continue the study of such processes as are of interest to him.

The book is clearly printed with easily read diagrams and clear photographs, is well-indexed for ready reference, and should make a valuable addition to any reference library.

R. R. Batcher†

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† Decatur Manufactury Co., Inc., Brooklyn, N. Y.

MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE*

THIS is a list of references prepared by the Bureau of Standards for the months of May and June, 1928. It is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The articles listed below are not obtainable from the Government. The various periodicals can be consulted at large public libraries or at The Engineering Societies Library, 33 West 39th Street, New York City.

R100. Radio Principles

R110 Lambert, A. Sur la vitesse de propagation des ondes radiotelegraphiques. (On the velocity of propagation of radio waves.) *Comptes Rendus*, **186**, pp. 686–688; March 12, 1928. *Experimental Wireless* (London), **5**, p. 285; May, 1928.

R113 Mesny, R. Observations et travaux recents sur la propagation des ondes electromagnetiques. (Recent observations and work on the propagation of electromagnetic waves.) *L'Onde Electrique*, **7**, pp. 129–155; April, 1928.

(Review of work done for two years on the question of propagation).

R113 Taylor, A. H. and Young, L. C. Studies of high-frequency radio wave propagation. Proc. I. R. E., **16**, pp. 561–578; May, 1928.

(Studies of multiple signals. Time of propagation of round-the-world signals and also of echoes having a very much shorter transmission distance. Method worked out for predicting in advance when round-the-world signals are likely to appear).

R113.3 Pratt, H. Apparent night variations with crossed-coil radio beacons. Proc. I. R. E., **16**, pp. 652–57; May, 1928.

(Describes night direction shifts and fading of signals from directive type of radio beacon as received on an airplane in flight.)

R113.3 Friis, H. T. Oscillographic observations on the direction of propagation and fading of short waves. Proc. I. R. E., **16**, pp. 658–665; May, 1928.

(Observations made on the 16-meter transatlantic signals to determine variation of direction of propagation and amount of fading.)

R113.5 Nodon, A. Relations entre la propagation des ondes electromagnetiques, l'activité solaire et l'état atmospherique. (Relations between the propagation of electromagnetic waves, solar activity and amount of atmospherics.) *L'Onde Electrique*, **7**, pp. 136–161; April, 1928.

(Values for speed of propagation of radio waves deduced from other data always leads to figures less than 300,000 km per second.)

R113.6 Munro, G. H. The reflecting layer of the upper atmosphere. *Experimental Wireless* (London), **5**, pp. 242–44; May, 1928.

(Experiments carried on in New Zealand during Dec., 1925 for estimation of height of reflecting layer for waves of 500 kc.)

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* Original Manuscripts Received by the Institute, May 21, 1928 and June 16, 1928.

R120 Green, E. Short-wave aerial systems—An elementary theory of the transmission of high-frequency energy along the feeders. *Experimental Wireless* (London), **5**, pp. 304–311; June, 1928.

(Discussion based on theory of Heaviside. Results of transmission along feeder lines under different conditions.)

R125.1 Aicardi, J. Sur un nouveau dispositif d'alignement par emissions Hertziennes. (On a new method of course setting by wireless.) *Comptes Rendus*, **186**, pp. 305–307; January 30, 1928. *Experimental Wireless* (London), **5**, p. 288; May, 1928.

(Uses two antennas short distance apart radiating same radio-frequency waves from one antenna pure CW and from other modulated. Course indication is produced by interference pattern).

R127 The equivalent inductance and capacity of an aerial. *Experimental Wireless* (London), **5**, pp. 297–299; June, 1928.

(Calculation of inductance and capacity of antenna by considering it as a transmission line.)

R130 von Ardenne, M., and Stoff, W. Zusammenfassender Bericht—Über die Kompensation der schadlichen Kapazitaten und ihrer Rückwirkungen bei Elektronenröhren. (Summarized report—On the compensation of the properties capacities and their reaction in electron tubes.) *Zeits. für Hochfrequenztechnik*, **31**, pp. 122–128, April; pp. 152–157, May, 1928.

(Theoretical discussion.)

R130 Ashworth, M. H. Graphical computations. *Experimental Wireless* (London), **5**, pp. 311–314; June, 1928.

(Applies graphical method of integration to electron tubes).

R131 Doring, E. Über den Einfluss hoher ohmscher Anodenwiderstande auf den Gitterwiderstand von Verstarkerröhren. (Influence of higher plate resistance on the grid resistance of amplifier tubes.) *Zeits. für Hochfrequenztechnik*, **31**, pp. 116–120; April, 1928.

(Experimental data on the mutual effects of the two internal resistances.)

R131 Lange, H. Die Stromverteilung in Dreielektrodenröhren und ihre Bedeutung für die Messung der Voltaspannung. (The current distribution in three-electrode tubes and their importance for the measuring of voltages.) *Zeits. für Hochfrequenztechnik*, **31**, pp. 105–109, April; pp. 133–140, May, 1928.

(Theoretical discussion of tube characteristics).

R132 Baggally, W. On banks of parallel valves feeding resistance loads without distorting the wave form. *Experimental Wireless* (London), **5**, pp. 315–321; June, 1928.

(Mathematical theory of design of bank of electron tubes for supplying power to loudspeakers. Treatment confined to case of a non-reactive load.)

R134 Ballantine, S. Detection by grid rectification with the high-vacuum triode. PROC. I. R. E., **16**, pp. 593–613; May, 1928.

(Method for experimentally ascertaining frequency distortion in detection. Method of securing efficient grid rectification in superheterodyne systems.)

R141 Bird, L. T. Reactance and admittance curves applied to tuned circuits with and without resistance. *Experimental Wireless* (London), **5**, pp. 327–334; June, 1928.

(Reactance and admittance curves developed from considering vector diagrams and properties of inductance and capacity.)

R143 Reed, M. Damping due to grid current in the case of a valve oscillator. *Experimental Wireless* (London), **5**, pp. 322–324; June, 1928.

(Calculation of damping introduced in an oscillatory circuit by grid current.)

R145.5 Allen, R. G. The establishment of formulae for the self-inductance of single turn circuits of various shapes. *Experimental Wireless* (London), **5**, pp. 259–63; May, 1928.

(Development of formulas.)

R152 Busse, E. Über eine Methode zur Erzeugung von sehr kurzen elektrischen Wellen mittels Hochfrequenzfunken. (On a method of production of very short electric waves by means of high-frequency sparks.) *Zeits. für Hochfrequenztechnik*, **31**, pp. 97–105; April, 1928.

(Production and measurement of characteristics.)

R200. Radio Measurements and Standardization

R201.6 Zickner, G. A bridge for the measurement of inductance and capacity. *Experimental Wireless* (London), **5**, pp. 280–282; May, 1928.

(Application of the Maxwell Bridge to measurements made by radio experimenters.)

R201.7 Schneider, W. A. Use of an oscillograph for recording vacuum-tube characteristics. Proc. I. R. E., **16**, pp. 674–80; May, 1928.

(Results obtained with oscillograph for plotting photographically vacuum-tube characteristics.)

R210 Dellinger, J. H. The status of frequency standardization. Proc. I. R. E., **16**, pp. 579–92; May, 1928.

(Accuracy of frequency standards. Intercomparison of frequency standards of U. S. with foreign countries reported.)

R210 Dellinger, J. H. Finding the "yardsticks" of radio at the Bureau of Standards. *Popular Radio*, **13**, pp. 290–91; April, 1928.

(Work of Radio Section of Bureau of Standards in frequency standardization.)

R210 Pierce, G. W. Magnetostriction oscillators. *Proc.* American Academy of Arts and Sciences, **63**, pp. 1–47; April, 1928.

(Description of method of using magnetostriction in connection with electron-tube circuits to produce and control frequency of electrical and mechanical oscillations. Range of frequencies from few hundred cycles per second to three hundred thousand cycles per second.)

R210 Black, K. C. A dynamic study of magnetostriction. *Proc.* American Academy of Arts and Sciences, **63**, pp. 49–66; April, 1928.

(Measurements of dynamic electrical characteristics of magnitude of vibration of various magnetostrictive rods and tubes.)

R213 Namba, Y. and Namba, S. Study on the operation of the multivibrator. Researches of the Electrotechnical Laboratory No. 218, November, 1927. Tokyo, Japan.

(Experimental and mathematical discussion of multivibrator.)

R214 Jouaust, R. Quelques modes particuliers de vibration des quartz piezoelectriques. (Certain modes of vibration of piezo-electric quartz plates.) *L'Onde Electrique*, **7**, pp. 125–128; March, 1928.

(Shows deformations of quartz plates which are more complex than have been found before.)

R220 Roberts, W. Van B. A capacity measurement method. *Jnl.* Franklin Institute, **205**, pp. 699–701; May, 1928.

(Method of calibrating condensers at radio frequencies.)

R230 Coil calculations—Design data for short wave coils. *Wireless World and Radio Rev.*, **22**, pp. 466–68; May 2, 1928.

(Charts giving calculations for coils in the 40–110 and 110–240 meter bands.)

R240 Jackson, W. Dielectric losses in single layer coils at radio frequencies. *Experimental Wireless* (London), **5**, pp. 255–258; May, 1928.

(Experiments to measure the increase of effective resistance of coils at radio frequencies.

R251.2 Spot welded thermojunctions. *Experimental Wireless* (London), **5**, p. 314; June, 1928.

(Description of instruments.)

R270 Judson, E. B. An automatic recorder for measuring the strength of radio signals and atmospheric disturbances. Proc. I. R. E., **16**, pp. 666–670; May, 1928.

(Apparatus described for automatically recording field intensity of low-frequency signals and atmospheric disturbances.)

R270 Sreenivasan, K. A short survey of some methods of radio signal measurement (concluded from April issue). *Experimental Wireless* (London), **5**, pp. 273–78; May, 1928

(Description of methods of measurement of field intensity used in various laboratories.)

R275 Robinson, E. H. The power in a modulated oscillation. *Experimental Wireless* (London), **5**, pp. 252–254; May, 1928.

(Interpretation of voltmeter and ammeter readings in terms of power dissipated in case of circuits carrying modulated radio-frequency currents.)

R300. Radio Apparatus and Equipment

R320.8 Smith-Rose, R. L. Wireless masts and screening. *Wireless World and Radio Rev.*, **22**, pp. 460–462; May 2, 1928.

(Shielding action of metal masts used at Rubgy station.)

R334 Decaux, B. Applications nouvelles des lamps a quatre electrodes. (New applications of four-electrode electron tubes.) *L'Onde Electrique*, **7**, pp. 119–124; March, 1928.

(Uses of these electron tubes for relays, multivibrators, etc. Plate on electron tube can be small and so the signal itself can be used to operate the tube.)

R334 Screened grid valves—Informal discussion at meeting of Wireless Section, Institution of Electrical Engineers of London. *Experimental Wireless* (London), **5**, pp. 335–38; June, 1928.

(Application of these tubes to radio-frequency amplification.)

R342 Kafka, H. Zur Niederfrequenzverstärkung mit Drosselspulenkopplung. (On the low-frequency amplification with impedance coupling.) *Zeitschrift für Hochfrequenztechnik*, **31**, pp. 87–90; March, 1928.

(How to design impedance-coupled amplifiers.)

R342.6 Thomas, H. A. Retro-action in amplifiers. *Experimental Wireless* (London), **5**, pp. 245–251; May, 1928.

(Gives general properties of regeneration and analysis of conditions to be fulfilled for advantageous use in amplifiers.)

R342.6 Glasgow, R. S. Tuned radio-frequency amplifiers. *Jnl.* A. I. E. E., **47**, pp. 327–331; May, 1928.

(Simplified method of calculating performance of tuned-circuit amplifying set with curves showing performance obtained with various types of tubes.)

R343.7 A German H.T. mains unit with glow discharge rectifier. *Experimental Wireless* (London), **5**, p. 251; May, 1928.

(A receiving tube for use on alternating current.)

R344 Über die gleichzeitige Erregung zweier Schwingungen in einer Dreielektrodenröhre. (On the simultaneous excitation of two oscillations in a three-electrode tube.) *Zeitschrift für Hochfrequenztechnik*, **31**, pp. 72–84; March, 1928.

(Discussion of possibilities of generating two oscillations within the same tube. Methods and use.)

R344.3 Byrnes, I. F. Recent developments in low-power and broadcasting transmitters. Proc. I. R. E., **16**, pp. 614–651; May, 1928.

(Description of various types of radio-transmitting equipment ranging in output from 200 to 2000 watts. Application of master-oscillator, power-amplifier circuits for low- and medium-frequency transmitting sets. Uses of piezo-oscillator control for high-frequency and broadcasting transmitting sets. Brief explanation given of equi-signal system of radio-beacon transmission.)

R351 Lampkin, G. F. How to build a beat-frequency oscillator. *Radio Broadcast*, **13**, pp. 156–158; July, 1928.

(Construction details of laboratory-type oscillator are given.)

R384.1 Griffiths, W. H. F. Substandard wavemeter design. *Experimental Wireless* (London), 5, pp. 324–26; June, 1928.

(Design of a radio-frequency frequency meter with new type of inductances for 30,000 kc.)

R384.1 Griffiths, W. H. F. The demonstration of a new precision wavemeter condenser. *Experimental Wireless* (London), 5, pp. 278–279; May, 1928.

(Sullivan-Griffiths variable air condenser for use as part of a substandard wavemeter).

R385.5 Jakowleff, A. J. Analyse einer Elektrischen Schaltung für des Kondensatormikrophon. (Analysis of an electrical connection for the condenser microphone.) *Zeitschrift für Hochfrequenztechnik*, 31, pp. 85–87; March, 1928.

(Theory of the condenser microphone.)

R388 Rangachari, T. S. The harmonic comparison of radio frequencies by the cathode-ray oscillograph. *Experimental Wireless* (London), 5, pp. 264–267; May, 1928.

(Use of cathode-ray tube for comparison of harmonics of radio frequencies.)

R400. Radio Communication Systems

R402 Kruse, R. S. Getting started at 30 megacycles. *QST*, 12, pp. 9–10; May, 1928.

(Description and design of receiving and transmitting apparatus for 10-meter work.)

R412 Blackwell, O. B. Transatlantic telephony—The technical problem. *Bell System Technical Jnl.*, 7, pp. 168–186; April, 1928.

(Non-technical description of engineering problems involved in developing transatlantic radio circuits by means of which the American Telephone and Telegraph Company's system is used for communication with England.)

R412 Waterson, K. W. Transatlantic telephony—service and operating features. *Bell System Technical Jnl.*, 7, pp. 187–194; April, 1928.

(Description of differences in operating practice on the two sides of the Atlantic. Data given on the extent to which transatlantic connection was used during first year.)

R500. Applications of Radio

R520 Donisthorpe, H. de A. Air service and amateur cooperation. *Wireless World and Radio Rev.*, 22, pp. 491–492; May 9, 1928.

(Report of coast-to-coast flight of all American airplane equipped with radio apparatus and assistance rendered by amateurs.)

R500. Applications of Radio

R525 Eisner, F., Fassbender, H., Kurlbaum, G. Leistungs- und Strahlungsmessungen an Flugzeug- und Bodenstationen. (Energy and radiation measurements on aircraft and ground stations.) *Zeits. für Hochfrequenztechnik*, 31, pp. 109–116, April; pp. 141–151, May, 1928.

(Measurements of power in antennas on aircraft and ground stations. T-shaped antenna used at ground station and trailing wire antenna on aircraft.)

R526.2 Franck, P. Le radiocompass et la navigation aerienne. (The radio compass and aerial navigation.) *L'Onde Electrique*, 7, pp. 109–118; March, 1928.

(Discusses an ideal radiocompass which could be used on aircraft for direct reading of bearings.)

R536 Eve, A. S. and Keys, D. A. Geophysical prospecting. *Scientific American*, 138, pp. 414–417, May; pp. 508–511, June, 1928.

(Description of various methods used. Based on Bureau of Mines Tech. Paper No. 420).

R550 Eckersley, P. P. The design and distribution of wireless broadcasting stations for a national service. ***Jnl.*** I. E. E. (London), **66**, pp. 501–528; May, 1928.

(Discusses problems in the distribution and design of broadcasting stations.)

R582 Clarkson, R. P. What hope for real television? *Radio Broadcast*, **13**, pp. 125–128; July, 1928.

(Resumé of the different methods of television.)

R800. Non-Radio Subjects

512 Ratcliffe, J. A. Symbolical algebra. *Experimental Wireless* (London), **5**, pp. 239–242; May, 1928.

(Discussion of errors which occur in applying process of symbolical algebra to treatment of problems containing sine and cosine functions.)

621.3 Buckley, O. E. High-speed ocean-cable telegraphy. *Bell System Technical Jnl.*, **7**, pp. 225–267; April, 1928.

(Brief history of development of permalloy-loaded cables and discussion of problems concerned with their design, construction, and operation.)

621.374.2 Herman, J. Bridge for measuring small time intervals. *Bell System Technical Jnl.*, **7**, pp. 343–349; April, 1928.

(Measurement of time intervals from about one ten-thousandths of a second up to several seconds is described.)

621.382 Pages, A. La telegraphie multiple par courantes de frequences audibles. (Multiplex telegraphy for currents of audible frequencies.)

(Description of cable system used by French company which uses band filters in cable telegraphy so that speed of signal will not be impeded by deformations which occur at certain frequencies.)

APPLICATIONS FOR MEMBERSHIP

Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Committee on Admissions. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before September 2, 1928. These applications will be considered by the Board of Direction of the Institute at its September 5th meeting.

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	New York City, 195 Broadway	Blackwell, O. B.
	New York City, 64 Broad Street	Kroger, Fred H.
	New York City, R. C. A., 66 Broad Street	Ranger, Richard H.
	New York City, General Electric Co., 120 Broadway	Young, Owen D.
	Schenectady, General Electric Co.	Baker, W. R. G.

For Transfer to the Member grade

Illinois	Chicago, 2022 The Engineering Bldg.	Turner, George S.
Maryland	Baltimore, Room 13, Custom House	Sterling, George Edward
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	Cambridge, 10 Wyman Road	Weeks, Paul T.
New Jersey	Newark, 200 Mt. Pleasant Avenue	Rypinski, Maurice C.
New York	Grasmere, S. I.	Grimes, David
	Great Neck, L. I.	Rigney, Douglas
	New York City, 40 Broad Street	Blenheim, William J.
	New York City, R. C. A., 70 Van Cortlandt Park So.	Engel, Francis H.
	New York City, 420 Lexington Avenue	Melhuish, Harold T.
	New York City, R. C. A., 66 Broad Street	Winterbottom, W. A.
Pennsylvania	East Pittsburgh, 124 Avenue A	Sutherlin, Lee

For Election to the Member grade

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	Rochester, 25 Leighton Avenue	Haugh, A. T.
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Peru	Lima, c/o Marconi's Wireless Telegraph Co., Ltd.	Madgwick, G.

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	Palo Alto, 804 Bryant Street	Dow, Clifford J.
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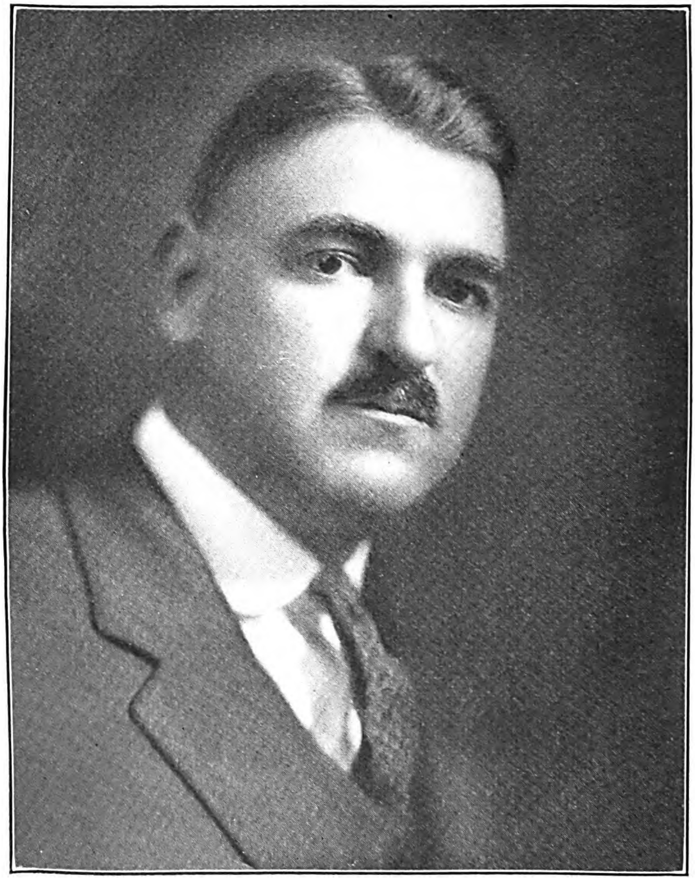

Leonard F. Fuller
Manager of the Institute, 1928

Leonard F. Fuller

MEMBER OF THE BOARD OF DIRECTION OF THE INSTITUTE, 1928

Leonard F. Fuller, born in Portland, Oregon, August 21, 1890, was graduated from Cornell University in 1912 with the M.E. degree. Immediately thereafter he entered the employ of the National Signalling Company of Brooklyn, N. Y. A few months later he joined the staff of the Federal Telegraph Company at San Francisco and in 1913 was made chief electrical engineer of that Company.

The ensuing six years were very actively filled with the development and manufacture of Poulsen arc transmitters for the Army and Navy. A number of stations for trans-oceanic service were built for the Navy in the United States, France, Panama and across the Pacific to the Philippines. While this work was under way Dr. Fuller contributed several papers on radio subjects to the PROCEEDINGS of the Institute. He also continued his studies, taking graduate work at Stanford University, from which he received the Ph.D. degree in 1919. This same year he was awarded the Morris Liebmann Memorial Prize by the Institute for his contributions to long-distance radio communication.

After the need for arc stations ended with the war, Dr. Fuller left the Federal Telegraph Company in 1919 to organize the Colin B. Kennedy Company of San Francisco for the manufacture of radio receivers. He was engaged in the activities of this company and in private consulting practice, specializing in power company communication problems, until 1923.

The next three years were spent in Schenectady and New York City in power company communication and radio receiver work. In 1926, Dr. Fuller returned to San Francisco for the General Electric Company in connection with new high voltage developments and the application of vacuum tubes to the problems of the light and power industry on the Pacific Coast.

Dr. Fuller was appointed a member of the Board of Direction of the Institute in January of 1928. He is Chairman of the San Francisco Section of the Institute, is a Fellow of both the Institute of Radio Engineers and the A.I.E.E., and a member of the American Physical Society.

CONTRIBUTORS TO THIS ISSUE

Austin, L. W.: (See PROCEEDINGS for February, 1928).

Ballantine, Stuart: (See PROCEEDINGS for April, 1928).

Breit, G.: Born in Russia, July 14, 1899. Radio Section, Bureau of Standards, 1918–1919, 1920–1921; received the Ph. D. degree from Johns Hopkins University in 1921; national research fellow, Leiden and Harvard Universities, 1921–1923; assistant professor, University of Minnesota, 1923–24; mathematical physicist, Department of Terrestrial Magnetism, Carnegie Institution of Washington since 1924.

Brown, J. E.: Born at East Marion, New York, September 11, 1902. Attended school of engineering, Cornell University, 1920–24. Test Department of General Electric Company, 1923; Radio Inspector with the U. S. Department of Commerce at Detroit, Michigan, 1924 to present date, engaged in inspection work, development of field-strength measuring apparatus, methods and measurements, and frequency measuring equipment. Member of the Institute.

Dahl, Odd.: (See PROCEEDINGS for March, 1928).

Diamond, Harry: Born February 12, 1900 in Quincy, Massachusetts. Received the B.S. degree from Massachusetts Institute of Technology in 1922 and the M.S. degree from Lehigh University in 1925. Engaged in air compressor work with the General Electric Company for one and one-half years; for four years instructor in electrical engineering at Lehigh University; summer work with the General Electric Company, the B.F. Sturtevant Company and the Boston Elevated Company. In Radio Section, Bureau of Standards, July 1927 to date. Associate member of the Institute.

Edwards, S. W.: Born at Dennison, Ohio, August 11, 1889. Employed by United Wireless Corporation of America 1908 to 1913; long-distance automatic calling, Ohio Telephone Company, 1913–14; appraisal engineering work, Columbus Railway, Light and Power Company, 1914–17; in charge of Eighth Radio District, U. S. Department of Commerce since 1917. At present time engaged as Supervisor of Radio, Eighth Radio District, in development of field-strength measuring apparatus and methods, constant frequency equipment and secondary standards for use by the Division, design of radio test cars for the Radio Division. Member of the Institute.

Hooper, Stanford C.: Associated with the radio work of the Navy since its earliest inception. Served two tours of duty as Fleet Radio Officer and three tours of duty as Head of the Radio Division of the Bureau of Engineering, Navy Department. During the Arms Conference in Washington served on the Advisory Committee; was also technical advisor to the American Delegation at the International Radiotelegraph Conference in Washington, 1927. Captain Hooper is at present Director of Naval Communications, with additional duties as Technical Advisor to the Federal Radio Commission. He is a Fellow in the Institute.

Shangraw, Clayton C.: Born at Richford, Vermont, April 2, 1896. Wentworth Institute, 1915–17; communications officer, Army Air Service, 1918–22; airplane pilot, New York Telephone Company, 1923; assistant and associate radio engineer, Signal Service at Large, Signal Corps Aircraft Radio Laboratory, Wright Field, Dayton, Ohio, 1927 to date. Associate member of the Institute.

Stowell, E. Z.: Born August 30, 1900 in Arlington, Mass. Received the B.S. degree from Tufts College, 1921; Ph. D. degree, American University, Washington, D. C., 1927. Assistant in physics at Tufts College, 1918–21; graduate assistant at University of Nebraska, 1921–23; With American Radio and Research Corporation in vacuum-tube research, 1920–21; assistant physicist, Bureau of Standards, June to August, 1922; In Optics Division, Bureau of Standards, 1924 to March 1925; assistant physicist, Radio Section, Bureau of Standards, March 1925 to December 1927. With the Federal Telegraph Company at Palo Alto, California, December 1927 to date.

Tuve, M. A.: (See PROCEEDINGS for June, 1928).

INSTITUTE ACTIVITIES

SUSPENSION OF BOARD MEETINGS DURING JULY AND AUGUST

No meetings of the Board of Direction of the Institute were held during the months of July and August. The Board meets again on September 5th.

NEW MEMBERS TO BE ELECTED

The attention of the Institute membership is called to the list of applicants for the various grades of membership, either for transfer or direct election, appearing on page 1270 of this issue. Applications from all of these members and non-members have been acted upon by the Committee on Admissions. These applications will be presented to the Board of Direction at its October 3rd meeting. Members objecting to the transfer or election of any persons listed should communicate with the Secretary of the Institute on or before September 30th.

1928 MEMBERSHIP CARDS

As announced on several occasions in this section of the PROCEEDINGS, membership cards are being sent only upon specific request this year. If there are any members who desire 1928 cards, a request therefor should be addressed to the Institute. Cards for all grades of membership are available.

1928 YEAR BOOK

The 1928 Year Book, mailed as a supplement to the July issue of the PROCEEEDINGS, is now in the hands of all members. Any member who failed to receive a copy can secure a duplicate free of charge from the office of the Institute.

Through typographical error there is an unfortunate mistake in connection with the listing of the subcommittees of the Committee on Standardization, to be found on page 20 of the Year Book, in that the Subcommittee on Receiving Sets, J. H. Dellinger, Chairman, was omitted. This subcommittee (as reported from time to time in this section of the PROCEEDINGS) has been very active and has made considerable progress which has contributed to the thoroughness of the Preliminary Draft of Report of the Committee on Standardization which is being considered by the entire committee. Since Dr. Dellinger assumed the chairmanship of the Committee on Meetings and Papers, E. T. Dickey has been chairman of the Subcommittee on Receiving Sets.

It will be appreciated if the membership will call to the attention of the Institute office any errors of commission or omission in the membership lists of the Year Book, not previously reported.

NEW BUREAU OF STANDARDS PUBLICATION

Beginning July, 1928, the Bureau of Standards publishes a new monthly journal, "Bureau of Standards Journal of Research", which will contain the bureau's research papers in the fields of physics, chemistry, engineering, and a number of special technologies including radio communication. The annual subscription rate is $2.75 with an extra charge of $0.75 for postage to countries outside the United States, Canada, Mexico and Cuba.

Members of the Institute will undoubtedly be interested in this journal. Subscriptions should be forwarded to the Superintendent of Documents, Government Printing Office, Washington, D. C.

1928–1931 CRUISE OF THE *CARNEGIE*

Terrestrial Magnetism and Atmospheric Electricity for March, 1928 contains a program of scientific work on cruise VII of the non-magnetic ship *Carnegie*, which is making a three-year tour of the world under the Department of Terrestrial Magnetism of the Carnegie Institution of Washington. A complete account of the work to be done, including a number of important investigations in the field of radio communication, appears in this issue of the above journal.

NEW STYLE OF I.R.E. EMBLEM

The new member's emblem in the form of a screw-back lapel button, approximately one-half the size of the present emblem, is now available in the four colors for the various grades of membership.

The new emblem is of 14 k. gold and can be purchased for $2.75 postpaid from the Secretary of the Institute.

Committee Work

COMMITTEE ON CONSTITUTION AND LAWS

The Committee on Constitution and Laws, R. H. Marriott, Chairman, has held a number of meetings throughout the summer in connection with its work in revising the Constitution and By-Laws of the Institute. The immediate work on the new Constitution has been completed by the Committee. The proposed Constitution will be submitted to the Board of Direction at its September meeting with a view to its being submitted to the membership of the Institute for approval at the same time the ballots for 1929 officers are mailed.

Changes in the Constitution have been slight, but the By-Laws of the Institute have been very extensively modified and amplified by the Committee.

COMMITTEE ON ADMISSIONS

The Committee on Admissions held its regular monthly meeting on August 1st. The following members were present: R. A. Heising, Chairman; F. H. Kroger, and E. R. Shute. The Committee acted upon 11 applications for transfer or election to the higher grades of membership in addition to the election of a large number of Junior and Associate members.

SUBCOMMITTEE ON BIBLIOGRAPHY

The Subcommittee on Bibliography of the Committee on Standardization held a meeting in the offices of the Institute on August 1st. The following members were present: C. A. Wright, Chairman; I. G. Maloff, W. C. White, C. E. Brigham, G. C. Southworth, R. D. Brown, Jr., and H. A. Frederick.

The principal questions under discussion at this meeting were (a) whether or not a bibliography should properly form a part of an Institute Standardization Report, (b) what should be the type and quality level of the references included, and (c) what should be the degree of completeness of the bibliography.

The opinion of the committeee was that the object of this bibliography is to amplify and make more intelligible the Standardization Report. Its educational object is warranted by the needs of those who will use the report. In view of this object it was agreed that the bibliography should contain a limited number of outstanding articles, some of them describing methods of testing and others presenting fundamentals of the various subjects covered by the Standardization Report. Recent articles will be given greater emphasis than older ones.

The kind and quality level of the articles included will be such that they will be intelligible to the average radio engineer throughout the country who might have use for the Standardization Report. Short descriptive paragraphs will be added after each reference, including information as to the kind and length of article. The references will be numbered in accordance with the Dewey System.

Personal Mention

C. D. Pitts is in the engineering department of the Electrical Research Products, Inc., at Philadelphia.

Burton E. Ebert has removed from Chicago to Philadelphia, where he is now connected with the Pooley Company.

E. H. Hansen of the Fox Case Movietone Corporation has been transferred from New York to Hollywood, California.

L. M. Cockaday, for a number of years technical editor of *Popular Radio,* is now on the radio staff of the New York Herald-Tribune.

Rear Admiral S. S. Robison, U. S. Navy, until recently Commandant of the 13th Naval District, is now Superintendent of the United States Naval Academy at Annapolis.

John B. Hawkins has recently become associated with the Newcombe Hawley Company of St. Charles, Illinois. Mr. Hawkins was formerly with the Nathaniel Baldwin Company.

Captain S. C. Hooper of the U. S. Navy is now Director of Naval Communications. He is also serving as one of the technical advisors to the Federal Radio Commission.

Charles H. Cross has resigned from the radio interference department of the Pacific Gas and Electric Company to become associated with the Radio Department of the Dollar Steamship Lines at San Francisco.

Thornton P. Dewhirst is practicing consulting work at 4909 Seventh Street, N. W., Washington, D. C. Mr. Dewhirst for several years was associated with the C. Francis Jenkins Laboratories of Washington.

J. H. Dellinger has been given a three months' leave of absence from the Radio Section of the Bureau of Standards to be Chief Engineer to the Federal Radio Commission. During Dr. Dellinger's absence from the Bureau, C. B. Jolliffe is serving as acting chief of the radio section.

OBITUARY

With deepest regret the Institute announces the death of

Frederick G. Simpson

For many years Mr. Simpson was connected with the Kilbourne-Clark Corporation of Seattle in the manufacture of radio apparatus. He served with that company as chief engineer, manager, and vice-president.

During the world war he served as radio material officer in the Navy at Boston, returning to civil life with the rank of Commander in the Reserve.

During more recent years he was directing engineer and vice-president of the Simpson Radio Corporation, which manufactured radio apparatus embodying Mr. Simpson's inventions. At the time of his death, Mr. Simpson was in charge of radio engineering for the Robert Dollar Steamship Company.

He was elected to Membership in the Institute in 1913 and was made a Fellow in 1915.

He died on March 16th, 1928 at Seattle, Washington.

Volume 16, Number 9 *September, 1928*

THE USE OF RADIO FIELD INTENSITIES AS A MEANS OF RATING THE OUTPUTS OF RADIO TRANSMITTERS*

By

S. W. Edwards[1] and J. E. Brown[2]

([1]Supervisor of Radio, U. S. Department of Commerce, Detroit, Michigan; [2] Radio Inspector, U. S. Department of Commerce, Detroit, Michigan.)

Summary—*A method is described by means of which the outputs of radio transmitters could be regulated by Federal authority as required by law in terms of measured radio field intensities instead of watts power in the transmitter or antenna circuits. The method was developed from surveys made on five different radio broadcast stations. For purposes of example, figures are given only for radio broadcast stations. The figures given may not be those which might be ultimately desirable for regulation purposes; however, they are usable as a means of explanation and demonstration and, therefore, serve the purpose for which this paper is intended. The method is developed on averages taken from actual measurements and is applicable to all conditions found in the broadcast band at the present time.*

In dealing with radio transmission as an efficient means of communication there are two factors vitally essential and under human control which should be regulated if maximum success is to be obtained. Without careful assignment and regulation of these two factors, complete satisfaction will not be realized generally. These factors are the power and frequency of transmission.

Since regulation of these two things is very necessary to successful radio transmission it would seem that each of them should be given careful study so that the best results possible may be achieved. The study of frequency regulation is quite complete and has been pursued closely since almost the very beginning of radio. Unfortunately, the study of power regulation has not received as much attention and as a result the same terminology and methods are used today as have always been used. With the hope that further interest will be stimulated in the subject this paper attempts to deal with the problem of power regulation of radio transmitters and sets forth a suggested method which is believed to be somewhat of an improvement over methods now used.

For purposes of example the regulation of power and the rating of transmitters operating in the broadcast band of 1500

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* Original Manuscript Received by the Institute, June 13, 1928.

to 550 kilocycles will be discussed, inasmuch as it seems that the greatest benefits of the method about to be outlined could be realized here, although the scheme is applicable to all frequencies with equal facility. Insofar as technical reasons are concerned it has been presumed that the power which could be alloted to a broadcast station was determined only by the amount of interference which would be caused another station operating on the same frequency, and by the amount of interference which the broadcast listener located near the transmitter would find in trying to listen to some other station on an adjacent frequency. These two reasons represent the technical premises on which power has heretofore been regulated. In considering these two things it should be borne in mind that neither of them were, or could have been, measured quantitatively by the regulating governmental bodies. The answers to both questions were determined largely by experiment, estimation, and past experience. From the time the first radio transmitter was built its power was expressed in terms of watts, and the external effect which it produced was estimated from this quantity insofar as regulation was concerned. The term watts of power has been widely interpreted and applied perhaps in the most popular forms of watts

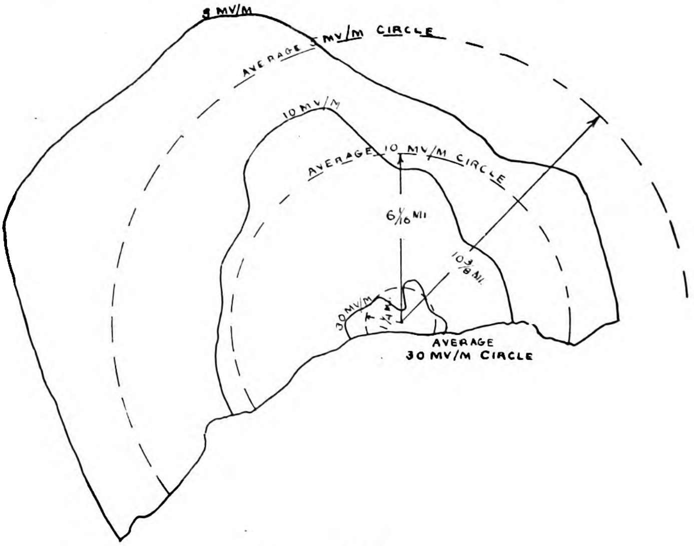

Fig. 1—Station A
Effective height, 21.7 meters; antenna current, 8 amperes; frequency, 850 kilocycles.

power in the antenna, or on the plates of the oscillator or radio-frequency amplifier tubes, or in terms of total watts input to transmitters. It is quite well recognized at this time that none of these methods, all of which in the end only show power somewhere in the circuit, is of any real value in determining the external effect which a radio transmitter will produce.

Considering the two technical premises on which the output of a radio transmitter is regulated, it will be readily seen that from the point of view of the present time and with present methods and equipment only one of these is of major importance. This is the regulation of power so that the broadcast listener does not get such a strong signal that with present day receiving equipment he cannot receive a reasonable number of stations without interference and still have a signal of maximum value for good high-quality reception. It is believed that it is quite generally conceded at this time that the operation of two or more radio stations on the same frequency is not good engineering practice unless they are of very low power and separated by very great distances. There are economic and technical reasons why this is so. When it is considered that a 1-kilowatt transmitter is classed as of very medium power at this time and when consideration is given to the cost of such a transmitter, it will at once be apparent that such transmitters should not be required to duplicate fre-

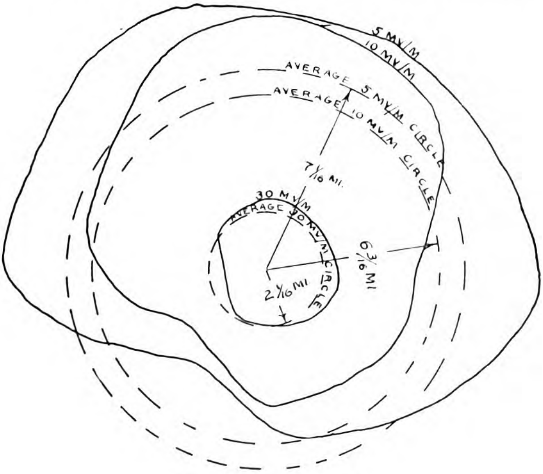

Fig. 2—Station B

Effective height, 39.5 meters; antenna current, 6 amperes; frequency, 700 kilocycles.

quencies, knowing as we do that under favorable conditions two such transmitters in the United States, irrespective of distance separation, will cause a beat note somewhere in this country. Considering then the regulation of power of a radio station in its most exact technical and economic sense, there is only one reason why power should be regulated, this reason being, as stated before, to see that the broadcast listener gets the best possible service without undue interference and that the broadcaster be placed in a position where he can realize the maximum result on his investment. On the basis of this argument, proposals and facts on the regulation of power of radio transmitting stations operating in the broadcast band, and with powers at one kilowatt and greater, are presented.

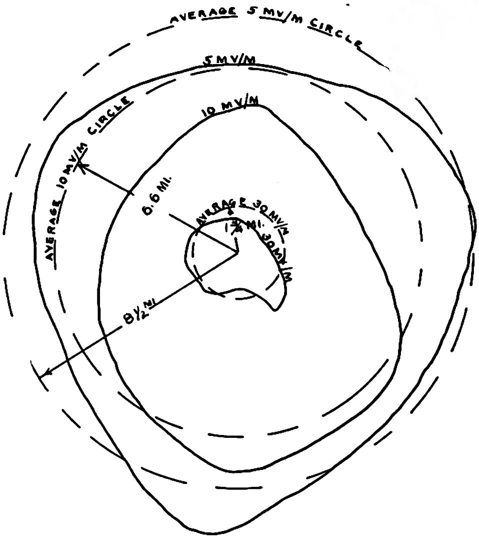

Fig. 3—Station C
Effective height, 23.5 meters; antenna current, 7.8 amperes; frequency 640 kilocycles.

A suggested method of regulating the outputs of radio broadcast transmitters not in terms of watts of power but in terms of measured radio field intensities is offered. These measured radio field intensities could be expressed in terms of microvolts or similar terms per meter, at specified distances. In short, it is suggested that the output of a radio transmitter be regulated on the basis of the actual external effect which it produces and not on the estimate of what effect a certain amount of power in watts found in some part of the transmitter will produce.

It has been quite definitely established by various capable and independent engineers that radio field intensities on the order of ten thousand microvolts per meter will furnish very excellent radio reception. Signals of this strength have a level above most interfering noises which includes ordinary summertime static and inductive disturbances. While signals below this value are capable of providing excellent reception, nevertheless, taken under all conditions of summer and winter and with all ordinary interferences, a signal intensity of ten thousand microvolts per meter is of such a level that high-quality reproduction will be obtained.

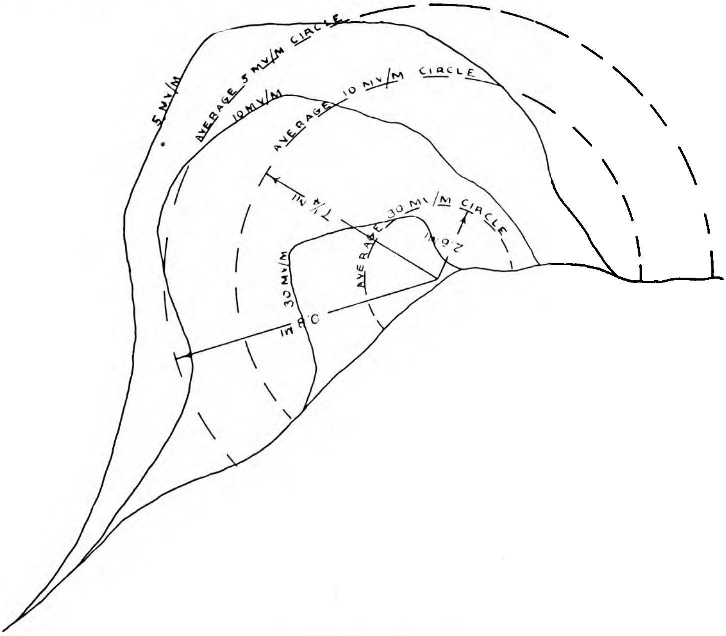

Fig. 4—Station D
Effective height, 32.3 meters; antenna current, 8 amperes; frequency, 770 kilocycles.

The problem then is to regulate the output of existing and future radio transmitters so that the largest possible number of broadcast listerners will receive signals of 10,000 microvolts per meter or more, at the same time bearing in mind that too great a signal must not be delivered to a large number of listeners. The matter is one of balance having the reception of a high quality signal on one side and not so great a signal as to interfere with reception on the opposite side.

For the purposes of this paper it was decided to call signals above 30,000 microvolts per meter interfering signals, or signals which would very likely not permit a broadcast listener with an average set to receive any considerable number of stations other than those with a level of 30,000 or more microvolts per meter. In order to have some flexibility on the question of just what constitutes the minimum signal strength which gives consistent high-quality reception, values down to 5,000 microvolts per meter are used. Keeping these three values as working figures and bearing in mind that in the ideal allocation of frequencies in the broadcast band the only reason for the regulation of power is for the purpose of giving the greatest number of broadcast listeners signals lying between the values of 5,000 and 30,000 microvolts per meter, the following method of power regulation was developed.

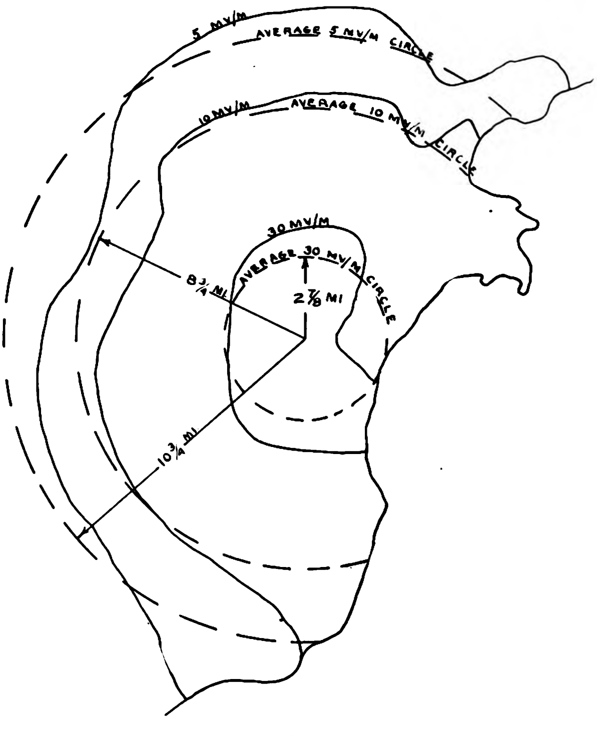

Fig. 5—Station E
Effective height, 18.8 meters; antenna current, 6.8 amperes; frequency 1110 kilocycles.

The figures and working data were taken from five radio field-intensity surveys made on the same number of different stations in various widely separated cities in the United States, east of the Mississippi River. The stations measured are typical of the trend of the art at the present time. (See Radio Service Bulletin, March, 1927). The stations surveyed had antenna input powers varying from 750 to 3,500 watts and were operating on frequencies from 640 to 1,070 kilocycles. Since the original surveys covered widely varying powers, it was necessary to reduce them all to the same value so that comparisons could be made. The arbitrary figure of 1 kilowatt was chosen as being typical of the average station and being a good value to work with, for the reason that a station of such power if properly located is capable of propagating a 10,000-microvolt-per-meter signal over the major portion of most cities. This reduction was accomplished by means of the conventional square law method, the results of which when checked in several ways against an actual reduction of power at the transmitter showed no error of any consequence. The operation involved the reduction of three powers, 750 watts, 1,500 watts and 3,500 watts to 1,000 watts on which power the other two stations measured were operating.

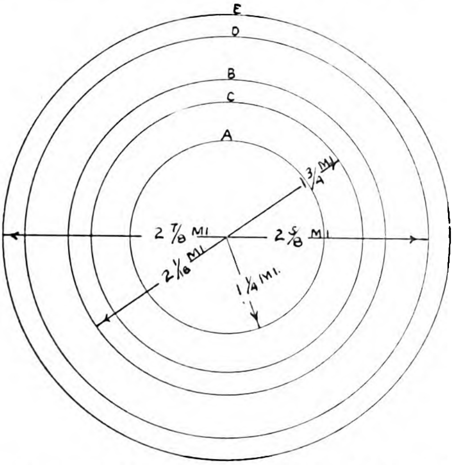

Fig. 6—Average 30-Millivolts-per-Meter Fields of Five 1,000-Watt Radio Stations.

The 5,000-, 10,000- and 30,000-microvolt-per-meter fields of each of the stations was plotted using the same scale dimensions. This operation then reduced the original measured values to the same comparable form in graphic manner. The five radio field-intensity maps placed in this form are shown in Figs. 1, 2, 3, 4, and 5.

Radio stations A, B, C, and D are located in cities varying in size from 500,000 to 1,500,000 population. Station E is located in the country some twenty miles removed from a city of 1,500,000 inhabitants. Since stations A, B, C, and D were located in all cases in metropolitan areas, in some cases directly on large steel buildings, in others adjacent to them, the fields are necessarily much more erratic and of considerably lesser extent than station E which is far removed from any absorptive or shadowing body. For this reason station E could not be in all cases compared with the other stations even after its field had been reduced to what would be its normal coverage at 1,000 watts of power.

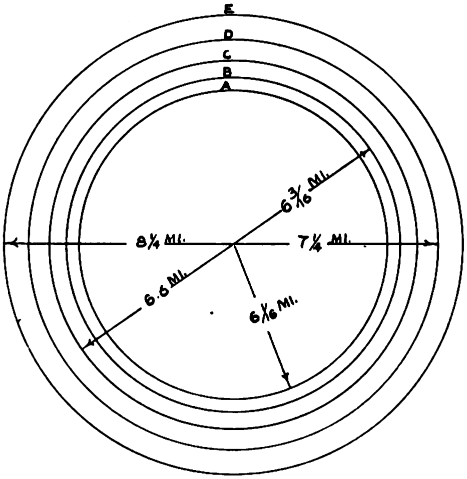

Fig. 7—Average 10-Millivolts-per-Meter Fields of Five 1,000-Watt Radio Stations.

Having reduced the surveys to that point where they represented equal values of output power at the transmitter and to the same scale of distances, the question arose as to what was the average distance at which the average 1,000-watt transmitter delivered fields of 5,000, 10,000 and 30,000 microvolts per meter. In searching for the answer the average 5,000-, 10,000- and 30,000-microvolt-per-meter fields of each of the stations were drawn. These are depicted by the circles shown on each of the field-intensity contour maps. This average field was arrived at for each contour line value by taking the average of the distances at which each individual contour occurred. Since the original surveys were made by making frequent measurements on a number of radii extending away from the transmitter as a center, the distances on each of these radii (at which for example the 30,000 microvolt per meter contour line was found) were averaged, giving as a result the average distance at which a 30,000 microvolt-per-meter signal might be found. This procedure was followed for each of the three stations giving as an individual result for each field the average distances at which signals of 5,000, 10,000 and 30,000 microvolts per meter are found. Circular fields for each value using the average distances obtained as radii are drawn on each of the contour maps. These average fields for the five stations are grouped under the three values of 5,000, 10,000, and 30,000 microvolts per meter and shown in Figs. 6, 7, and 8.

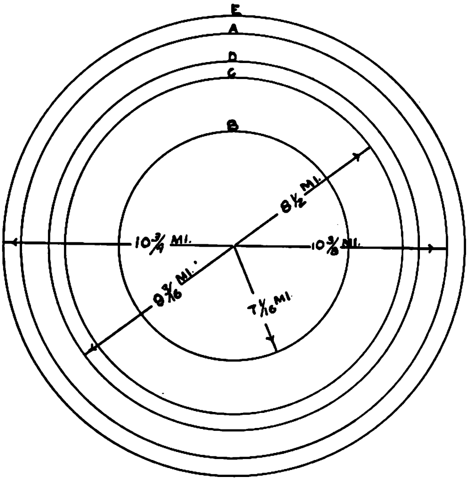

Fig. 8—Average 5-Millivolts-per-Meter Fields of Five 1,000-Watt Radio Stations.

The average of all of the distances at which for example the 30,000-microvolt-per-meter field occurred was then taken. This value shows the average distance at which will be found the 30,000-microvolt-per-meter signal from a 1,000-watt transmitter. The same procedure was followed for the 5,000- and 10,000-microvolt-per-meter lines. The average distance for the 30,000-microvolt-per-meter signal is found to be 2⅛ miles; for the 10,000-microvolt-per-meter signal 6$^{15}/_{16}$ miles or for all practical purposes 7 miles, and for the 5,000-microvolt-per-meter signal 9¼ miles. These three average fields are shown in Fig. 9. These values of signal voltage and of distance are of course all based on stations located in cities of some size and, therefore, are rather heavily attenuated. Computed radii of the ideal average fields of these four stations operating under conditions of no attenuation would show that a field of 30,000 microvolts per meter might be expected at four miles, a signal voltage of 10,000 microvolts per meter might be expected at 12 miles, and a signal voltage of 5,000 microvolts per meter might be expected at a distance of 24 miles. Estimated values of the average distances at which these three signal levels might be expected on a 1,000-watt station located in the country would be 30,000 microvolts per meter at 3½ miles, 10,000 microvolts per meter at 11 miles, and 5,000 microvolts per meter at 22 miles. These last mentioned figures are based on the results of the one survey made on a well-located station in the country.

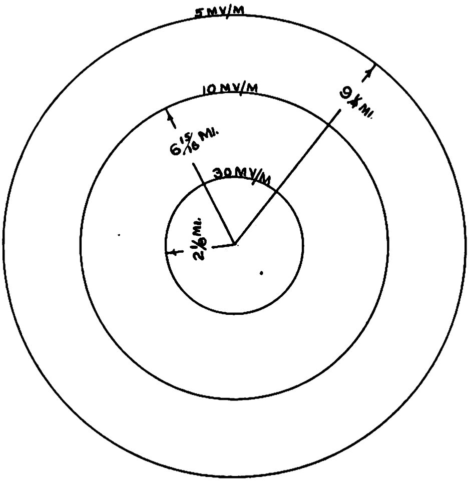

Fig. 9—Average 5-, 10-, and 30-Millivolts-per-Meter Fields of A Typical 1,000-Watt Radio Station.

Of all these figures those derived for the three values of fields of a station located in the city are at the present time most interesting and perhaps most important. They at once afford an avenue of approach to the much-debated questions of just how much power a station located in the business or other district of a city should use and just where the station should be located.

Although these figures represent an average value, the greatest deviation of any one of them from the average in the specific case of 10,000 microvolts per meter at 7 miles is $^{15}/_{16}$ mile or 13 per cent. In the case of stations located in the city, this can be considered a very good figure when all of the factors which enter into radio transmission under such conditions are considered.

The figures show quite conclusively that most 1,000-watt transmitters located in business districts in cities produce practically the same external field. By virtue of this fact, then, the value of 10,000 microvolts per meter at 7 miles becomes a real working figure for 1,000-watt transmitters, and there is no reason why this figure cannot be made to serve as a standard from which the field intensities at various distances for other powers can be computed. In connection with this standard figure the other two values of 30,000 microvolts per meter at 2⅛ miles and 5,000 microvolts per meter at 9¼ may also be classed as of similar value and equally applicable.

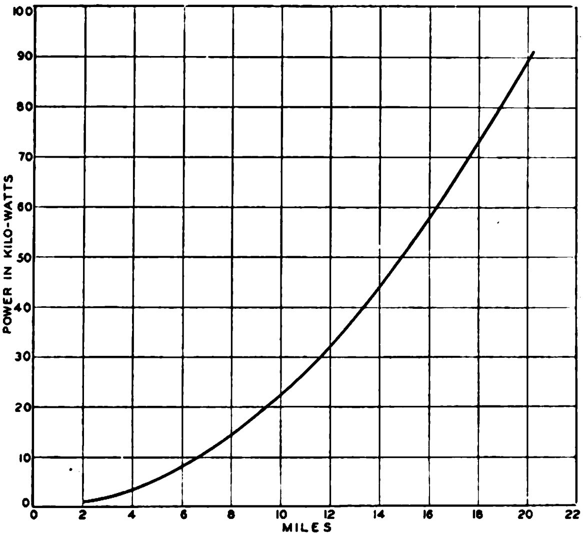

Fig. 10—30-Millivolts-per-Meter Field at Various Distances with Various Powers with City Attenuation.

The greatest deviation from the 30,000-microvolt-per-meter field at 2⅛ miles is 40 per cent. While this is large insofar as percentage is concerned when it is applied to the possible deviation it becomes small since the deviation is not greater than 0.8 of a mile. This special instance of a 40 per cent variation is due to the fact that one of the stations had a very small 30,000-microvolt-per-meter average field, and therefore becomes of still less importance. But for this small field the deviation would not be greater than approximately 25 per cent, which is really the fair percentage to be applied.

The value of 5,000 microvolts per meter at 9¼ miles likewise does not vary from the individual average curves by more than 22 per cent. While these figures are taken from surveys of only five different stations, the results obtained seem to indicate that there is a very considerable similarity between them, and there is reason to believe that this similarity holds for the majority of radio stations in operation.

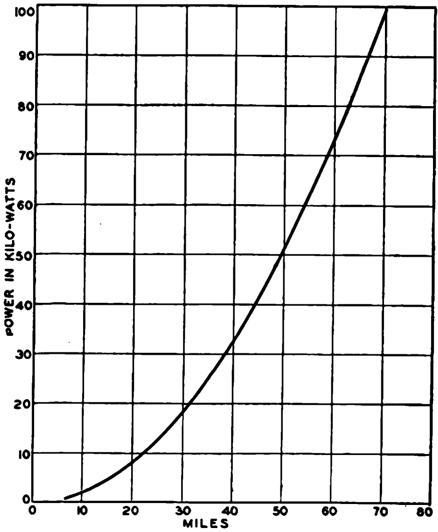

Fig. 11—10 Millivolts-per-Meter Field at Various Distances with Various Powers with City Attenuation.

Using these results as a working basis, the following method is suggested by which the output of any radio station can be regulated in terms of signal strength. The examples chosen are typical of every case of radio transmission which a governmental regulating body has to consider.

As a first case, an individual or organization decides that it would be desirable to enter the broadcasting field. The law requires that application forms be submitted for a radio station construction permit before any part of the station can be constructed.

This form shows exactly where the station will be located, the facilities which will be used, the type and power of transmitter, as well as the reasons for which it is desired. If the station is located in the residential section of a town or city, it is quite obvious that it cannot be permitted to set up a field in excess of 30,000 microvolts per meter over an area in which a large number of listeners reside. If there are many listeners within the 2⅛-mile circle, it is quite obvious that the power must be less than 1,000 watts. If the station is in the center of a business section the confines of which extend 2⅛ miles approximately in all directions from the transmitter, the power can certainly be of at least 1,000 watts, and should the business district be larger the power can be increased accordingly. It will be seen at once that there is a limit to this flexible condition in all instances with but one or two exceptions, these exceptions being the two largest cities in this country which make it doubtful if the power of city-located stations can ever exceed perhaps 2,000 to 3,000 watts without causing undue interference. The limit of the two exceptions mentioned is perhaps reached between 3,000 and 5,000 watts.

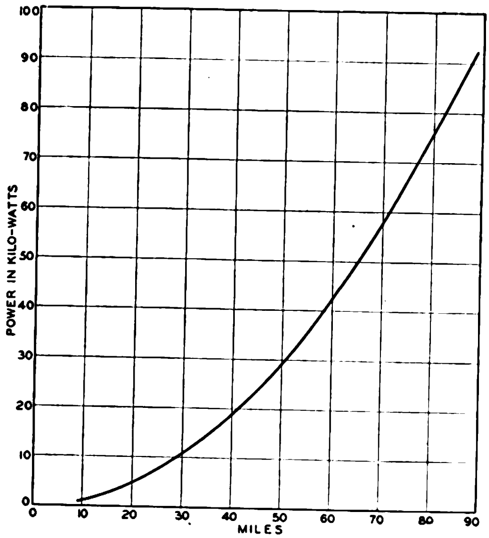

Fig. 12—5-Millivolts-per-Meter Field at Various Distances with Various Powers with City Attenuation.

Considering these things, then, anyone putting in a radio broadcast station in a city can so locate it that he can use as much as 1,000 or 1,500 watts successfully by carefully choosing the station location.

Now if it is desired at some future time to increase the station's power to perhaps 5,000 watts, it is necessary that a suitable site in the country be chosen for the station's location. Here again the station is located so that the 30,000-microvolt-per-meter signal does not have within its confines too great a population. Reference to the curves in Figs. 10 and 13 will show that for a 5,000-watt transmitter, a location between 4.5 and 9 miles away from the city limits can be chosen. Since the station is to be located in the country and away from all absorptive and shadowing bodies, it follows that the attenuation will be less, and therefore the signal will be somewhat greater than that predicted by the curve with attenuation. A value of distance of perhaps seven to nine miles would then seem advisable. These curves also show the distances which must be used for all powers up to 100,000 watts. Curves 2 and 3 show on what distances 5,000 and 10,000 microvolts per meter signal respectively can be expected.

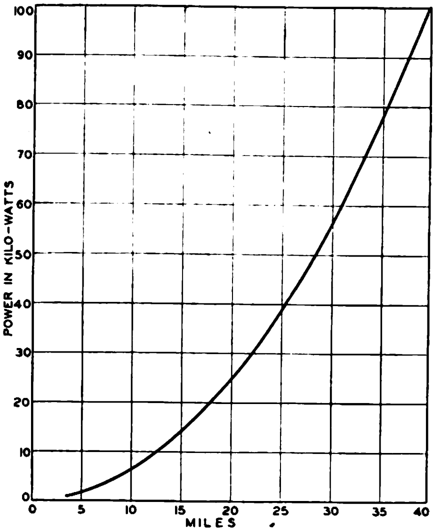

Fig. 13—30-Millivolts-per-Meter Field at Various Distances with Various Powers with No Attenuation.

Figs. 10, 11, and 12 are drawn under the conditions of average medium-sized city attenuation and are typical of the least distance at which signals of the values shown can be expected for given amounts of power. They represent the lower limit. Figs. 13, 14, and 15 represent the distances at which signals of 5,000, 10,000, and 30,000 microvolts per meter with no attenuation can be expected. They represent the maximum possible distances under any conditions over which signals of the values given could ever be expected. These signal values are given for all distances possible with powers up to 100,000 watts.

These two sets of curves under the two conditions of maximum and minimum attenuation are the limits of ordinary radio broadcast transmission and at a glance show what can be expected from a radio transmitter of any power in use in broadcasting today. They are, of course, just as applicable to any amount of power that the future may bring forth.

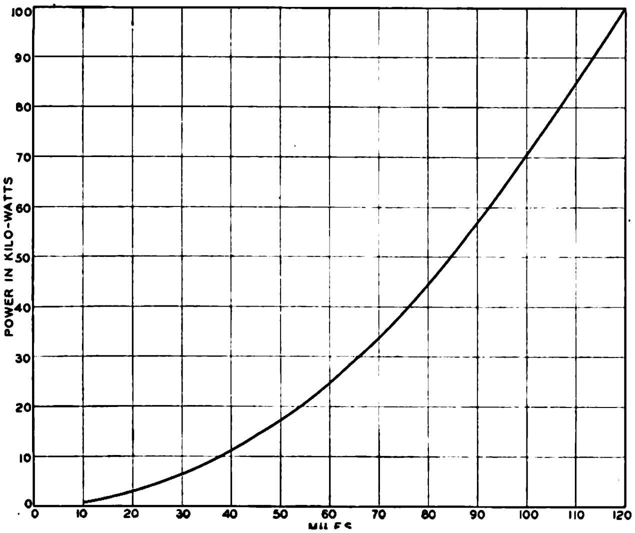

Fig. 14—10-Millivolts-per-Meter Field at Various Distances with Various Powers with No Attenuation.

To return to the case of the new broadcast station, the applicant may upon the advice of the regulatory body so locate this apparatus that the maximum amount of output can be used. This location has now been made on the basis of actual working figures and is no more the result of estimation and guess. Certainly this method is much better than any which has been used in the past or is now used. The applicant has only to say where he desires the good service area to be, and how large he desires it to be with respect to the location of his station and the maximum amount of power possible can be granted for this location. The benefits accruing to the broadcast listener are so apparent under this scheme as to require no discussion.

Now that the station's location has been decided upon and the power has been granted in accordance with the location, the construction permit can be granted. A field-strength survey is then made of the station using this power and the ultimate possible and final power allotment is decided upon from the results of the survey. The problem has then been given a solution and the station located and powered on the basis of sound engineering principles. The broadcaster has at once been placed in a position to realize the maximum on his investment and the broadcast listener has been given the conditions most favorable to best reception.

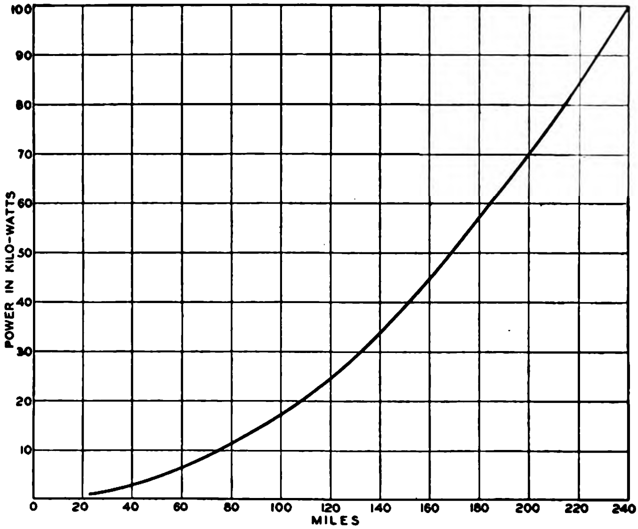

Fig. 15—5-Millivolts-per-Meter Field at Various Distances with Various Powers with No Attenuation.

Were the output of radio broadcast stations to be regulated on the basis of signal voltages instead of watts power somewhere in the circuit of the transmitter as at the present time, the greatest problem would be the disposition of existing stations. The answer would be found in identically the same manner as described in the case of the new station.

As an example a transmitter which is now using 1,000 watts of power could so continue provided that no great number of listeners reside within the $2^1/_8$-mile circle. If the station is so located that this is not the case, then the power of the station must be reduced so that the 30,000-microvolt-per-meter field does not encompass too many listeners. Should the listeners be located at a greater distance the power may be accordingly raised until the limit is reached.

Stations now classed as 100- or 500-watt transmitters, or any other power, would be regulated in exactly the same manner as described above for higher powers.

In each case, the method would be to assign the power which the charts included in this paper would indicate as being capable of producing the maximum possible field up to 30,000 microvolts per meter to the edge of heavy population in any locality. This amount is determined by reference to the values as shown on the charts having included attenuation and those without. At the first opportunity then an actual radio field-intensity survey on the station would be made and the final output determined as a result of this survey.

The third and final example of what would happen were the method of power regulation described to be adopted would be that of an existing station which desires to raise its power. The question would quickly be decided as in the two previous cases by reference to the proper charts. These would show just how much of an increase could be granted, keeping in mind the principles set forth in this paper as governing factors. If the station desired to raise power to such an extent that the necessity arose for a country location, the exact distance of the station from any great number of broadcast listeners could be at once stated, and the final allotment set by a regular survey of the field.

Should this new method of power regulation be adopted, no great difficulty would be experienced in introducing the change. In the first place, the Radio Division of the Department of Commerce through its field offices is well-informed of the conditions surrounding each of the broadcast stations now in existence. This necessitates a working knowledge of every one of the cities of importance in the United States with reference to just where the station is located, how far distant the majority or even any considerable number of broadcast listeners reside, and the other points which have immediate bearing on the subject. This information is possessed by the Radio Division and, therefore, makes this part of the problem comparatively simple. In any case the final value of power would be determined by a field-intensity survey on each station so that should any errors have been made in the preliminary assignment they would be corrected.

The question of just how the change from the old to the new method could be made has been met with a suggested answer. Another problem also arises as to just what under this new method the stipulations in the license should be. It seems quite proper to suggest that the principles already laid down be made to serve again and that the license be made to read that for the case of a given station its field shall not show an excess of 30,000 microvolts per meter at any distance greater than that at which a considerable number of broadcast listeners are found, this distance to be expressed in miles as found for the specific case. The approximate amount of power or antenna current which will set up this field can also be stated for the convenience and guidance of the station's operating personnel. For the specific case of a station which is located approximately 2⅛ miles from any great number of listeners, the license could be made to read perhaps in this fashion. "This station is permitted to so operate that at no distance greater than 2⅛ miles plus or minus 10 per cent shall there be found a radio field intensity greater than 30,000 microvolts per meter plus or minus 5 per cent."

The variation of 5 per cent plus or minus in the instances of field intensity and 10 per cent for distance would seem advisable so that the irregularities of the fields of stations may be taken care of satisfactorily. The figures 5 and 10 per cent are suggested ones and can finally be settled when a greater number of surveys have been made. If necessary as a further precaution, the distances at which the 5,000 and 10,000 microvolts per meter signals plus or minus 5 per cent variation in signal and 10 per cent in distance variation can also be set forth.

So far in this paper there has only been considered a new method of rating and regulating the output of radio broadcast stations and a method by which the new scheme could be put in force.

The method by means of which each of the surveys for this purpose would be made is described. As stated before, the tentative values of 5,000, 10,000, and 30,000 microvolts per meter signal strength are the important ones insofar as power regulation is concerned. The individual surveys on each station could, therefore, be made with a view toward the emphasis of the contour lines having these values.

The radial method of making surveys has been used exclusively in gathering the data used in this paper and has been found highly satisfactory.

The procedure followed in such a survey is to select streets or roads leading away from the transmitter radially making measurements at frequent points along each of these radial directions as far as is required. Past practice has been to make measurements every three-tenths of a mile, making each measurement at a street corner or other place which can be accurately found on a map of the locality under survey. By this means each measurement can be definitely and accurately located when the contour lines are drawn. The short distance between measurements is a reasonable insurance that no great change can take place in the signal being measured without becoming at once apparent. The radial directions are selected so that as the distance increases outward from the center or transmitter, the distance between points on any two adjacent radii does not become great. As a typical example, a station on which measurements were made for a distance of approximately 8 miles in each direction was measured along radii separated by angles of approximately 35·deg. Experience seems to indicate that the method is a sound one and that it is easily applicable under all conditions.

Following this method, readings will be made which will definitely locate the 30,000-,10,000-,and 5,000-microvolt-per-meter lines. By searching for only these three lines it becomes possible to make a fast survey containing all of the desired information. When gathering the data for the 30,000-microvolt-per-meter contour line, information as to the locations of any large number of broadcast listeners would also be collected and properly noted on the radio field-intensity map.

A brief analysis of the important factors about each of the stations might be of interest. Station A is located in a large city to the west of a number of large steel buildings, its antenna being hung between two steel buildings. Station B is located in a large steel structure north of the steel building area of a medium-sized city. The antenna system is rather poor, the ground system is only fair, and the soil conditions in the country immediately surrounding the transmitter are not conducive to best results.

Station C is located in a medium-sized city having no great number of steel buildings, none of any great size. A good antenna system is used and conditions are generally conducive to good transmission. Station D is located to the east of a large city having many steel buildings of large size. It is itself located on a large steel building. Station E is located 15 miles outside the city limits of a large city in a purely country location. Transmission from it is naturally very good.

There has been presented in the foregoing paragraphs a suggested method by means of which the rating of outputs of radio transmitters would be changed from terms of power in the transmitter to terms of signal strength. It is well-recognized that no method or principle, whether scientific or not, is of any value unless it is of a general nature. It must be such that all variable factors entering into the question can be used in a standard manner, which will always produce the same result. No principle can ever be a true one unless it contains within the statement of its make-up on applicable generalization of all the factors upon which it is founded. Every attempt has been made to subject the principles of power regulation on the basis of signal voltages as stated to as vigorous tests as could be found so that a true generalization might be drawn.

The undesirability of giving the power rating in watts of a radio transmitter as an index of its ability as a transmitter becomes more and more apparent. In these days of high-quality transmission, the term has long out-lived its period of usefulness. It is well-known that stations of equal power do not necessarily serve the same areas equally well. Many cases of this condition are evident today in radio broadcasting. This factor is of the utmost importance to the broadcaster and to the broadcast listener. And with the progression of time it is going to become increasingly so. The broadcast listener can never be assured of optimum reception until sufficient power is used in radio broadcasting to set up fields of a high order. The broadcaster who today stands the expenses of broadcasting can never be placed in his best position until existing stations radiate sufficiently strong signals to furnish reliable service over the areas desired. Certainly all of these facts and indications make essential the adoption of a method of power output regulation which is based on actual results secured and not on a figure which is comparatively meaningless. The only answer discernible at this time seems to be the measurement of radio field intensities and the use of these measurements for regulation purposes.

Experimental work over a period of two and a half years indicates that field-intensity measurements have many meritorious features. It is extremely unfortunate that more work in this field has not been done and that methods and apparatus have not been earlier developed. It is only of comparative recent date that accurate equipment has been made available which can be used for this work. It is earnestly hoped that with the advent of this apparatus the study of radio transmission phenomena will receive more interest and will continue on a wider scale than in the past.

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NOTE ON RADIO-FREQUENCY TRANSFORMER THEORY*

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Summary*—In the mathematical analysis for radio-frequency amplifiers employing tuned transformer coupling, the effect of the distributed capacitive coupling existing between the transformer windings has heretofore been neglected. General equations including this effect are here developed. It is shown that with these equations a closer agreement between computed and experimental results can be obtained.*

The presence of capacitive coupling changes the amount of mutual inductance necessary to obtain optimum amplification at a given frequency, at the same time reducing the optimum voltage gain. The manner of variation with frequeny of the equivalent transformer primary impedance is also altered.

RADIO-FREQUENCY transformer theory has for some time been reduced to mathematical treatment.[1] The accepted equations are simple and usually quite effective. In many cases, however, computations based on these equations are not in entire agreement with careful laboratory measurements. As will be shown in this note, this lack of agreement is due chiefly to the existence of a capacitive coupling between the primary and secondary windings which modifies the transformer performance, but the effect of which is considered negligible in the usual analysis. When this factor is not neglected, the resultant equations, which will be developed below, yield a closer agreement with experimental data.

An exact consideration of the capacitive coupling is difficult because of its distributed nature. In Fig. 1 is shown the equivalent circuit diagram corresponding to a stage of radio-frequency amplification employing tuned transformer coupling, in which the distributed capacitive coupling is replaced by a lumped value C_o. Since the primary and secondary windings are ordinarily connected together at one end through the batteries,

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Publication approved by Director, National Bureau of Standards of the U. S. Department of Commerce.

[1] Electric oscillations and electric waves. G. W. Pierce. McGraw-Hill Book Co.

A mathematical study of radio-frequency amplification. V. G. Smith. PROC. I. R. E., 15, 525–536; June, 1927.

Tuned radio-frequency amplifiers. R. S. Glasgow. *Journal* A.I.E.E., 47, 327–331; May, 1928.

this approximation should prove sufficiently close, particularly since the effective value of C_o can be readily measured.[2] R_1 represents the tube plate resistance plus the radio-frequency resistance of the transformer primary, L_1 the primary inductance, M the mutual inductance between windings, and R_2, L_2, and C_2 the constants of the secondary circuit.

With the currents as indicated, the circuit equations are, respectively:

$$E_1 = R_1 I_1 + j\omega L_1 I_2 - j\omega M I_3 \tag{1}$$

$$O = R_2 I_3 + j\omega L_2 I_3 - j\omega M I_2 - \frac{j}{\omega C_2}(I_1 - I_2 + I_3) \tag{2}$$

$$E_1 = R_1 I_1 - \frac{j}{\omega C_o}(I_1 - I_2) - \frac{j}{\omega C_2}(I_1 - I_2 + I_3) \tag{3}$$

The solution of the above equations for I_3 is of chief interest. An expression for the equivalent transformer impedance when referred to the primary side will also be derived. Substituting the value for I_2 from (1) in (2) and (3) and simplifying, (4) and (5) are obtained.

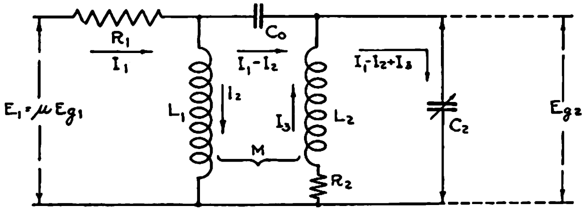

Fig. 1—Equivalent Circuit of A Stage of Radio-Frequency Amplification in Which Effect of Capacitive Coupling between Transformer Windings Is Considered.

$$I_1\left\{\frac{MR_1}{L_1} - \frac{R_1}{\omega^2 C_2 L_1} - \frac{j}{\omega C_2}\right\} + I_3\left\{R_2 + j\omega L_2 - \frac{j}{\omega C_2} + \frac{jM}{\omega C_2 L_1} - \frac{j\omega M^2}{L_1}\right\}$$
$$= E_1\left\{\frac{M}{L_1} - \frac{1}{\omega^2 C_2 L_1}\right\} \tag{4}$$

$$I_1\left\{R_1 - \frac{R_1}{\omega^2 C_2 L_1} - \frac{R_1}{\omega^2 C_o L_1} - \frac{j}{\omega C_2} - \frac{j}{\omega C_o}\right\}$$

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[2] This measurement was carried out by the usual negative intercept method of determining distributed capacity.

$$+I_3\left\{\frac{jM}{\omega C_2L_1}+\frac{jM}{\omega C_oL_1}-\frac{j}{\omega C_2}\right\}=E_1\left\{1-\frac{1}{\omega^2C_2L_1}-\frac{1}{\omega^2C_oL_1}\right\} \qquad (5)$$

Solving for I_3 and rationalizing

$$I_3=\frac{E_1\left\{\frac{M-L_1}{C_2}+\frac{M}{C_o}\right\}\left\{\left(\frac{Z_1}{C_2}+\frac{Z_2}{C_o}\right)-j\left(\frac{Z_3}{C_2}+\frac{Z_4}{C_o}\right)\right\}}{\left(\frac{Z_1}{C_2}+\frac{Z_2}{C_o}\right)^2+\left(\frac{Z_3}{C_2}+\frac{Z_4}{C_o}\right)^2} \qquad (6)$$

where

$$Z_1=R_1(L_1+L_2-2M)+R_2L_1-\frac{R_1}{\omega^2C_o}$$

$$Z_2=R_1L_2+R_2L_1-\omega^2R_1C_o(L_1L_2-M^2)$$

$$Z_3=\omega(L_1L_2-M^2)-\frac{R_1R_2+\frac{L_1}{C_o}}{\omega}$$

$$Z_4=\omega(L_1L_2-M^2+L_1R_1R_2C_o)-\frac{R_1R_2}{\omega}$$

Similarly, solving (4) and (5) for I_1 and placing $Z_1{}^1=\frac{E_1}{I_1}$, we have for the equivalent transformer primary impedance

$$Z_1{}^1-R_1=\frac{\{AC+BD\}+j\{DA-BC\}}{C^2+D^2} \qquad (7)$$

where $A=(L_1L_2-M^2)(C_2+C_o)-\frac{L_1}{\omega^2}$

$$B=\frac{R_2L_1}{\omega}(C_2+C_o)$$

$$C=R_2L_1C_2C_o-\frac{R_2}{\omega^2}(C_2+C_o)$$

$$D=\omega C_2C_o(L_1L_2-M^2)-\frac{1}{\omega}(C_2L_2+C_o[L_1+L_2-2M])+\frac{1}{\omega^3}$$

Equation (7) will be referred to later.

Referring to Fig. 1, the output voltage Eg_2 is equal to very nearly $\omega L_2 I_3$. The stage voltage amplification is consequently

$$K = \frac{Eg_2}{Eg_1} = \frac{\mu\omega L_2 I_3}{E_1} \tag{8}$$

where I_3 is as defined by (6). At a fixed frequency the voltage amplification is a maximum when I_3 is a maximum. Considering the tuning capacity C_2 as the variable, I_3 is a maximum when $\frac{dI_3}{dC_2}$ equals zero. Performing this operation two solutions are obtained both of which, upon eliminating negligible terms, resolve into (9).

$$\frac{Z_1}{C_2} + \frac{Z_2}{C_o} = 0$$

Substituting the values for Z_1 and Z_2 from (6) and simplifying, (10) is obtained for the adjustment of C_2 necessary to make I_3 a maximum. ($R_2 L_1$ has been dropped out, being negligible in comparison with the other terms).

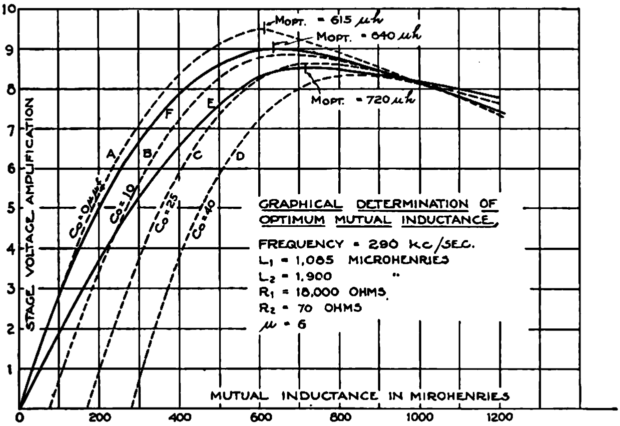

Fig. 2—Graphical Determination of Optimum Mutual Inductance

$$C_2 = \frac{\frac{1}{\omega^2} - C_o(L_1 + L_2 - 2M)}{L_2 - \omega^2 C_o(L_1 L_2 - M^2)} \tag{10}$$

Substituting the condition expressed by (9) in (6) and simplifying, we have (11), in which C_2 has the value indicated in (10).

$$I_{3_{\max}}=\frac{jE_1\{MC_2-(L_1-M)C_o\}}{\dfrac{L_1}{\omega}-\omega R_1R_2L_1C_2C_o-\left\{\omega(L_1L_2-M^2)-\dfrac{R_1R_2}{\omega}\right\}\left\{\{C_2+C_o\}\right.} \tag{11}$$

It is interesting to note that in the special case when no capacitive coupling exists ($C_o=0$), (10) and (11) reduce to

$$C_2=\frac{1}{\omega^2L_2} \tag{12}$$

$$I_{3_{\max}}=\frac{j\omega ME_1}{\omega^2M^2+R_1R_2} \tag{13}$$

which are, respectively, the value of C_2 necessary to resonate the secondary when acting alone and the maximum secondary current obtained by adjusting C_2 according to the simpler analysis (which neglects C_o).

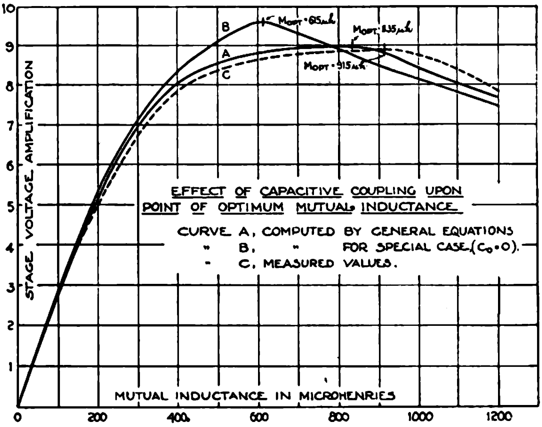

Fig. 3—Effect of Capacitive Coupling upon Point of Optimum Mutual Inductance.

Referring to (11) and (13), and considering M as the variable factor, $I_{3\max}$ will reach an optimum value when $\dfrac{dI_{3\max}}{dM}=0$. For the simpler case, when C_o is negligible, this yields (14) as the condition for optimum coupling at a given frequency.

$$\omega^2M^2=R_2R_1 \tag{14}$$

When the effect of capacitive coupling between windings can not be neglected, however, the condition for optimum amplification is given by the more complicated expression of (15) (the differentiation being based on the approximation that C_2 is independent of M).

$$M^2 - \frac{2L_1C_o}{C_2+C_o}M - \left(\frac{R_1R_2}{\omega^2} - L_1L_2 + \frac{L_1}{\omega^2(C_2+C_o)} - \frac{R_1R_2L_1C_2C_o}{C_2+C_o}\right) = 0 \quad (15)$$

Equations (14) and (15) are equivalent when C_o is equal to zero.

Since, in general, a variation in M involves also a variation in L_1 and C_o, it is evident that an accurate predetermination of the optimum mutual inductance is possible only when C_o is negligible.

It is practicable to keep L_1 constant while M is varied. C_o will also vary. The value of optimum inductive coupling may

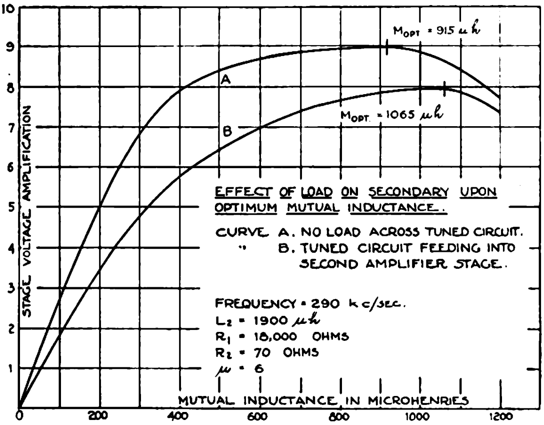

Fig. 4—Effect of Load on Secondary upon Optimum Mutual Inductance.

then be obtained by a graphical solution, as illustrated in Fig. 2. The constants of the experimental transformer used in this computation are collected in Table I. Curves A, B, C, and D were calculated by (8) and (11) and correspond to four assumed values of C_o. Since C_o increases as M is increased, the actual variation of amplification with mutual inductance will be as indicated by curve E. The ordinates of this curve will lie progressively on curves A, B, C, and D depending upon the value of C_o corresponding to a given value of M. Curve F shows how the amplification varies with the mutual inductance when the relative capacitive coupling is smaller than for curve *E*. Note that as the capacitive coupling is increased, the mutual inductance necessary for optimum amplification at a given frequency also increases, the actual magnitude of the optimum amplification being somewhat reduced.

TABLE I

CONSTANTS OF EXPERIMENTAL TRANSFORMER

Transformer Winding	Type of Winding	No. of turns and Size of Wire	Winding Length	Inductance in micro-henries	Resistance at 290 kc ohms
Primary	1¼ in. single layer solenoid	250 turns No. 34 single enameled wire	1¾ in.	1085	
Secondary	1¾ in. single layer solenoid	250 turns No. 34 single enameled wire	1¾ in.	1900	70

In the laboratory, *M* was varied by varying the number of turns on a 1¼ in. diameter primary coil form, using the secondary indicated in Table I. The mutual inductance was adjusted to the desired value and L_1, C_0 and *K* were measured. Data of a sample run are given in Table II. Using these data, curve *A* of Fig. 3 was calculated by means of (8) and (11). When the capacitive coupling is zero, (8) and (13) may be used; curve *B* corresponding to this condition. Curve *C* shows the actual variation of amplification with *M* as obtained by test. It will be observed that when the effect of C_o is considered, a much better agreement with laboratory measurements is obtained. Even then there is a slight discrepancy between computed and measured

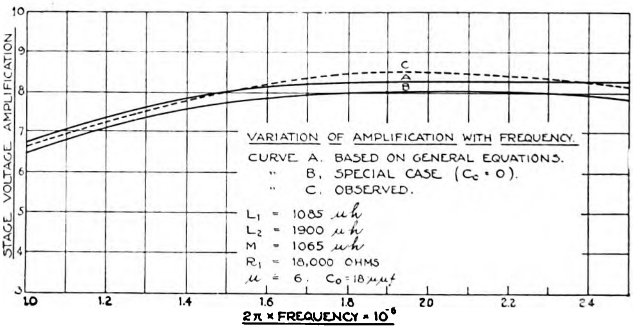

Fig. 5—Variation of Amplification with Frequency.

values; due in part to the effect of the tube output capacitance and partly to the increase in R_2 because of the tube voltmeter connected across the tuned circuit. Both effects tend towards overcoupling. The curves of Fig. 4 show that the point of optimum amplification occurs at a still larger value of mutual inductance if the stage output circuit feeds into a second amplifying stage, thereby further increasing the effective secondary resistance. Curve *A* corresponds to curve *C* of Fig. 3, while curve *B* is for the condition outlined, this being the usual state of affairs in practice. The optimum mutual inductance becomes 1065 rather than 915 microhenries.

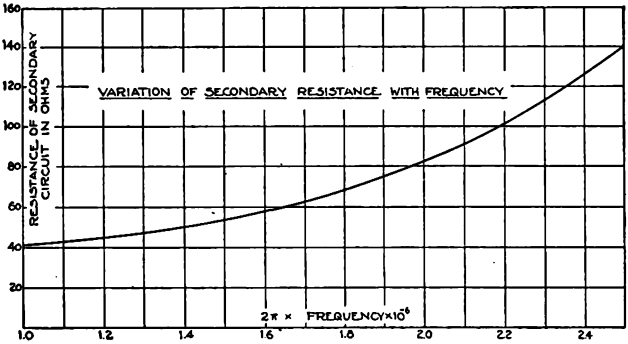

Fig. 6—Variation of Secondary Resistance with Frequency.

TABLE II

SAMPLE TEST DATA

Freq.	R_2 ohms	L_2 micro-henries	R_1 ohms	μ	M micro-henries	L_1 micro-henries	C_o $\mu\mu$f	Stage voltage amplification
290 kc per sec.	70	1900	18000	6	0	0	0	0
					200	103	8.1	5.05
					400	289	11.5	7.90
					600	514	14.2	8.65
					800	748	16.3	8.90
					1000	1000	18.2	8.80
					1092	1265	20.0	8.45

The manner of variation of stage voltage gain with the frequency is indicated in Fig. 5. Curve *A* is plotted from computations based on the general equations in which the effect of capacitive coupling is considered; curve *B* from computations based on the simpler equations resulting when $C_o = 0$, and curve *C* from actual measured results. Here, again, the agreement between computed and experimental values is better when the effect of capacitive coupling is not neglected. Fig. 6 shows how the resistance of the transformer secondary varies with frequency. This curve is essential in the computation of the curves of Figs. 5 and 7.

In conclusion, it is interesting to note that the equivalent primary impedance of the transformer is materially altered from that predicted by the simpler theory. According to the usual analysis, curves showing the variation of R_1^1 and X_1^1 with frequency are similar to those for a parallel resonance circuit. Actually, the variation indicated in Fig. 7 obtains, these curves being computed from (7). Within the useful frequency range of the transformer, there is no transition point at which X_1^1 changes from an inductive to a capacitive reactance, as is usually assumed.

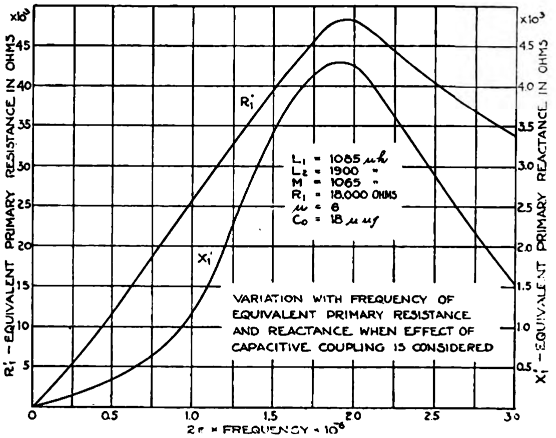

Fig. 7—Variation with Frequency of Equivalent Primary Resistance and Reactance When Effect of Capacitive Coupling Is Considered.

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RADIO BEACONS FOR TRANSPACIFIC FLIGHTS*

BY

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Summary—*This paper deals with the radio beacon and its practical application in connection with long distance flights over water. It describes briefly the operation of the equi-signal radio-beacon system as developed by the Air Corps and Signal Corps Aircraft Radio Laboratories at Dayton, Ohio, during the past few years.*[1]

The use of radio beacons as an aid to aerial navigation was considered necessary by the Army Air Corps in the plans for the flight from San Francisco to Hawaii by Lieutenant Hegenberger and Lieutenant Maitland. The Signal Corps was then requested to establish a radio-beacon station at each of the above points. The equipment for these installations was assembled and tested at the Signal Corps Aircraft Radio Laboratory and then shipped. One station was installed at Crissy Field, San Francisco and the other one was installed on the island of Maui, Territory of Hawaii. Several tests were made by these stations prior to the flight on June 28, 1927 during which the signals from each station were heard over the entire distance of twenty-four hundred miles.

The Army flight is described and the log kept by each station is shown. The beacon signals assisted materially in the problem of navigation during this flight, but unfortunately were not followed the entire distance due to difficulties with the receiving equipment. The next flight, over the same course, in which the radio beacons were used was that made by Messrs. Smith and Bronte. Here again trouble was experienced with the receiving equipment aboard the airplane which prevented the flyers from utilizing the beacon service continuously.

The Honolulu Chapter of the National Aeronautical Association requested the Army to provide radio-beacon service for the Dole Flight from San Francisco to Hawaii which was scheduled to start on August 12. This request was acceded to as it was desired to assist the participants as much as possible and further tests of the radio beacon were also desired. This flight was by far the most successful from a radio viewpoint, as the beacon signals were followed continuously by the winners of the Dole Flight. Photographs of the equipment are shown together with maps and the log kept by each station.

After the Dole Flight the beacon station on the Island of Maui was removed and permanently installed at Wheeler Field near Honolulu, where it is to be used in connection with local and inter-island flying by Air Corps personnel.

Since that time several improvements have been made in beacon equipment at the Signal Corps Aircraft Radio Laboratory which render it much more useful and reliable.

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* Original Manuscript Received by the Institute, June 13, 1928.

[1] Captain W. H. Murphy and Lieut. L. M. Wolfe, "The Stationary and Rotating Equi-signal Radio Beacons" *Journal* of Automotive Engineers, **19**, September 1926.

DURING the last year a keen interest in aviation has been aroused through the medium of long spectacular flights such as that of Lindbergh from New York to Paris, the Army flight from San Francisco to Hawaii by Lieutenant Hegenberger and Lieutenant Maitland, the Dole Derby from San Francisco to Hawaii, the Round-the-World flight by Brock and Schlee, and several others, some of which ended in disaster. The flight of the *Southern Cross* from San Francisco to Australia is perhaps the most vivid in the minds of the general public.

The airplanes used on these spectacular flights represented the last work in airplane design and equipment. Despite this, but few of the airplanes participating were equipped with radio apparatus of any kind. The omission of this seemingly important adjunct to navigation and safety, especially on any long flight over water, lies in the desire to carry a maximum fuel load, limiting all accessories to an absolute minimum. This attitude is quite natural, for it is impossible to calculate exactly how much fuel will be necessary for a flight, due to such factors as adverse weather conditions and other unforeseen reasons. With very few exceptions, if any, at the conclusion of these flights the pilots and navigators have expressed a belief that radio communication was essential and should be provided in future flights.

The preparations for the Army flight to Hawaii, made by Lieutenant Hegenberger and Lieutenant Maitland in a three-motored Fokker monoplane, did not include radio equipment until a few weeks before the proposed date of departure. Lieutenant Hegenberger, an expert aerial navigator and pilot, had made previous flights making use of the equi-signal radio beacon during its development at McCook Field, Dayton, Ohio, and realized its possibilities as an aid to aerial navigation, especially since the beacon equi-signal zone followed a Great Circle course. It was, therefore, recommended that a suitable radio-beacon station be installed at the starting point and also at the destination of the flight. The advisability of this plan was concurred in by the Chief of Air Corps, and the Chief Signal Officer was requested to have two radio-beacon installations made available for this project. The preparations for this project were carried out secretly so as to avoid any publicity before the proposed date of departure from San Francisco.

At this point a brief description of the equi-signal radio beacon will be given for those who are not familiar with the system. It is essentially a Bellini-Tosi system used for transmitting signals which are so arranged that they interlock and form a single dash along the bisector of the projected angle of the two inner coils, or primary, of the goniometer. The antenna consists of two triangular loops 80 feet in height and 300 feet along the base. These two loops must be very accurately laid out at right angles to each other and constructed so as to be identical in frequency and radio-frequency resistance. The goniometer is then connected at the base of the loops and equal values of capacity and inductance are

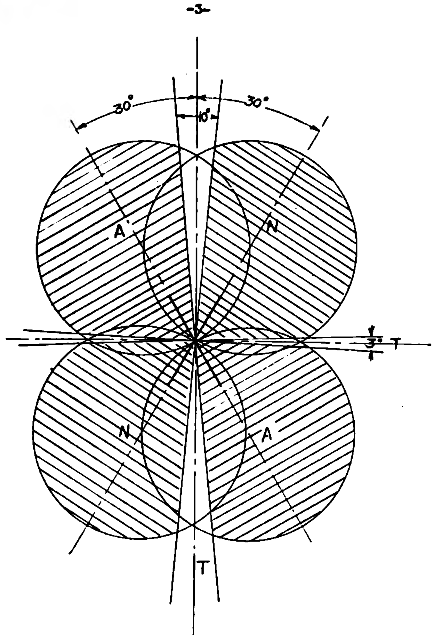


Fig. 1—Equi-Signal or *T* Zones of the Radio Beacon.

added in each leg of the loops so that the entire system is electrically balanced. The rotor coils of the goniometer are spaced 60 degrees apart, as this was found to be a fair compromise for obtaining a reasonable transmission range with minimum equi-signal zones narrow enough to allow one to follow the predetermined course with accuracy. See Fig. 1.

The primary coils of the goniometer are alternately connected to the output of the transmitter by means of an automatic relay that is actuated by letter *N* and *A* cams which are driven by a suitable electric motor. Thus, the letters *N* (–·) and

A (· –) in Morse code are used as a medium for breaking up the continuous dash obtained when the key of the transmitter is closed. The degree of coupling between these two primary coils and the loop coupling coils, or secondary, of the goniometer governs the current induced into each loop so that the letters *N* and *A* are radiated with varying intensity at different settings of the goniometer. When a receiver is so located as to pick up signals from these two loops with equal intensity the *N* and *A* signals interlock so as to form a single dash or *T*. This is equivalent to rotating two large loops with 60 degree displacement. Now, if the receiving set were located to one side or the other of the bisector of this projected angle then the signals would no longer be of equal intensity and the letter *N* or *A* would be heard, depending upon which side of the bisector the receiver was located.

In applying the above principles in practice the goniometer is set to permit an equi-signal or *T* to be transmitted along a predetermined course which is the line of flight along the Great Circle. As long as the airplane continues along this course the letter *T* is heard when the simple receiving set is tuned to the frequency of the beacon transmitter. If the airplane goes off the course slightly the *T* signal becomes broken, since the letter *N* is heard more loudly than the letter *A*. This has the effect of an *N* superimposed on a slightly weaker *A*. However, as the airplane gets further off the course in this direction the letter *N* becomes more distinct with the background character entirely lacking. Should the airplane get off the course to the left the letter *A* will be heard in a like manner.

In air navigation, as in marine navigation, it is necessary for the pilot to correct his course for drift due to side winds. This is accomplished by noting the position of ground objects through a drift indicator which is attached to the side of the airplane and then altering the course so as to compensate for the drift component. However, in flying above clouds or over water where no land marks are available the problem of correcting for wind drift becomes more difficult. In the transoceanic flights fixed ground objects were replaced by white-caps on the waves for reference in some cases, while in others smoke bombs were used. Any of these methods of checking the extent of wind drift and making the necessary changes in the compass course, whether of the magnetic or earth inductor type, require considerable skill and good judgment on the part of the navigator.

The equi-signal zone of the radio beacon, by virtue of its following the Great Circle course, automatically compensates for any directional error due to cross wind or to personal errors in making changes in the compass course in following the Great Circle. Thus, it will be seen that as long as the letter *T* is heard by the pilot or navigator the airplane is being held on the correct course, regardless of the wind drift. Whenever letter *T* becomes broken up so that a letter *N* is heard this indicates that the airplane is getting off the course to the right and when the letter *A* is heard it is an indication that the airplane is veering to the left of the course. This method, therefore, not only permits the pilot to keep a correct course, but also shows to a certain extent the amount of deviation from the course when the letter *N* or *A* is heard superimposed upon the background character. While the radio beacon indicates the desired course it is advisable to use the customary and time-honored compass so that the two methods may be used as a check against each other. In this case the radio-beacon signals need not be listened to continuously but periodically to check the compass course. In addition to the automatically-keyed range sign transmitted it is also possible to broadcast important meteorological data with the beacon transmitter.

THE ARMY FLIGHT

The plans for the use of the radio beacon for the Army flight from San Francisco to Hawaii called for an installation at Crissy Field, Presidio of San Francisco and a similar installation in the Hawaiian Islands. The advisability of this plan will be readily seen when it is shown that the entire Hawaiian Island group subtends but an angle of 7 deg. with San Francisco taken as the point of origin. This shows the necessity for setting and following a very accurate course toward the center of the island group, as an error of but 3½ deg. from this course would result in an airplane flying right on past the islands. With one beacon station located at San Francisco the initial course would be accurately established, but due to the widening of the equi-signal zone when going away from the station the accuracy of the course would be decreased. With a similar beacon located at the destination, however, the equi-signal zone would converge upon approaching the Hawaiian Islands, thus increasing the accuracy of the bearing sufficiently to bring the airplane directly towards the beacon station.

When the Signal Corps Aircraft Radio Laboratory was instructed during the latter part of March to install two beacons for the San Francisco-Hawaii flight a rather difficult problem was presented. The only beacon apparatus available was installed in the experimental station at Wilbur Wright Field. This equipment consisted of an obsolete 5 kw quenched spark transmitter and two experimental goniometers together with the necessary auxiliary apparatus for one beacon installation. The tentative date of the proposed Army flight was set as June 15, which allowed less than three months in which to obtain transmitting sets, construct and test additional apparatus, pack and make shipment to San Francisco and Hawaii, and then to install the apparatus at each place.

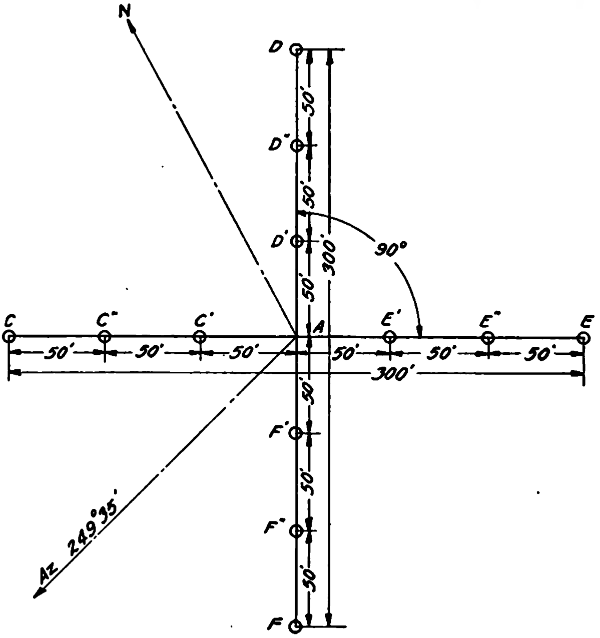
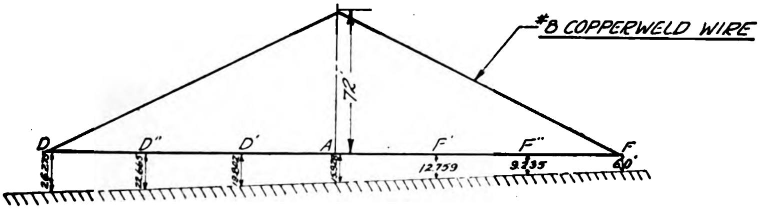


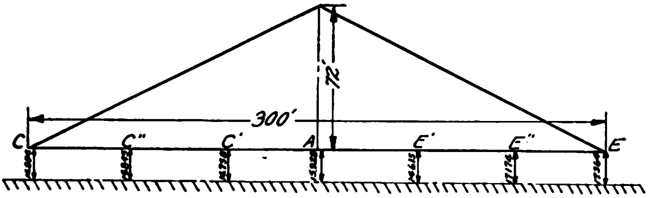
Fig. 2—Radio-Beacon Loop System, Crissy Field, California.

Attempts to purchase and get delivery on two 5-kw tube transmitters in this short period of time were futile, so it became necessary to modify existing 5-kw spark transmitters into tube transmitters. These sets utilized two 1-kw air-cooled tubes, type UV-851, in a full-wave self-rectified circuit with 500-cycle plate supply and operated on a frequency of 290 kc. The power equipment consisted of the 500-cycle motor generator sets which were previously used with the old quenched spark sets. Current for operating these machines was to be obtained from a standard Signal Corps d. c. generator driven by a gasoline engine. Complete equipment for the two radio-beacon installations was shipped by express to San Francisco on May 6, arriving in time for the equipment, consigned to Hawaii, to be loaded in an Army transport which sailed a week later.

Lieutenant Hegenberger, co-pilot and navigator of the proposed flight, having been previously stationed in the Territory of Hawaii, was familiar with the location and topography of the various islands, and upon his recommendation it was decided to locate the Hawaiian beacon station at Halawa Point on the eastern tip of the island of Molokai, or in the vicinity of Kahului, on the island of Maui. These locations come very close to the center of the Hawaiian island group. Since but a small variation in the course from San Francisco to Hawaii was permissible, it was considered advisable for the course to be set from San Francisco to the center of the group of islands and then, after sighting the islands, to change the course of the airplane to the island of Oahu, on which Wheeler Field is located.



LOOP DAF



LOOP CAE

Fig. 3—Radio-Beacon Loop System, Crissy Field, California.

The author arrived in San Francisco on May 12 and reported to the Signal Officer, Ninth Corps Area, for temporary duty. A suitable site for the San Francisco beacon was selected on the top of a hill at Crissy Field. The services of personnel of Company E, 5th Engineers, stationed at Ft. Scott, were made available for surveying and laying out the loop system. Credit is given this organization for the efficient and expeditious manner in which this work was carried out. The loop layout is shown in Figs. 2 and 3. A 90-foot pole for supporting the loops was purchased locally and arrangements made to have it set in place, and also to erect a temporary building in which to install the beacon equipment. A representative of the Signal Corps Aircraft Radio Laboratory arrived at Crissy Field on May 13, 1927 for duty in connection with the establishment of this beacon station and tests therewith.[2] Fig. 4 shows location of beacon station. Fig. 5 shows the interior of this station.

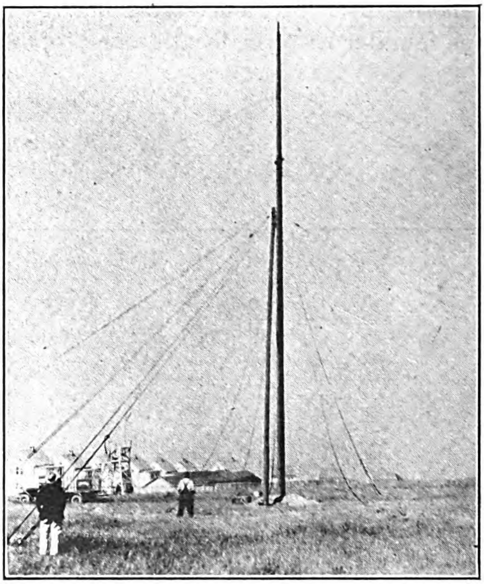


Fig. 4—Location of Radio-Beacon Station at Crissy Field; Pole in Position.

On May 18 the author proceeded via commercial liner to Honolulu, arriving on the morning of the 25th and reported to the Hawaiian Department for temporary duty in connection with the establishment and test of the radio beacon station in Hawaii.

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[2] Mr. L. A. Hendricks.

The Department Air Officer was requested to authorize a reconnaissance flight over the Island of Molokai and Maui for the purpose of selecting a suitable site for the establishment of the radio-beacon station. This flight, in which three planes took part, was made on May 27, 1927.[3] We flew over the Leper Colony on Molokai and over Halawa at a low altitude to learn as much as possible of its topography. It took but a short time to decide

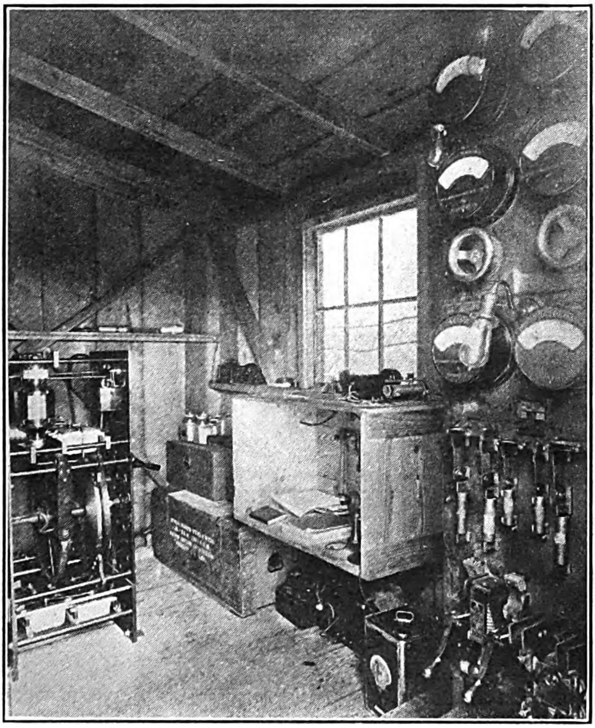

Fig. 5—Interior of Radio Beacon Station at Crissy Field.

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[3] Due to the hazard of inter-island flying with land type airplanes Army orders require that not less than three airplanes will participate in an inter-island flight. One of the airplanes is equipped with a radio transmitter and receiver and carries an expert radio operator who transmits the position of the airplanes to Headquarters every 5 minutes. In case of a forced landing in the water one of the airplanes goes back to Luke Field, near Honolulu, for assistance while the other circles above the disabled airplane.

against this place as a site for the beacon station due to its close proximity to rugged mountains and lack of docking facilities for unloading supplies and equipment. The shore line is extremely broken and edged with coral reefs, making it practically impossible to land even a small boat. We then proceeded across the channel toward Kahului on the Island of Maui. The windward side of this island appeared well-suited for a beacon location and we attempted to land at Kahului to look over the ground more carefully. The landing field at Kahului, however, was partially covered with water, making a landing difficult, so we proceeded to the Island of Lanai where the airplanes were refueled with Government gasoline and oil, held there for such contingencies, before the return to Luke Field, Oahu.

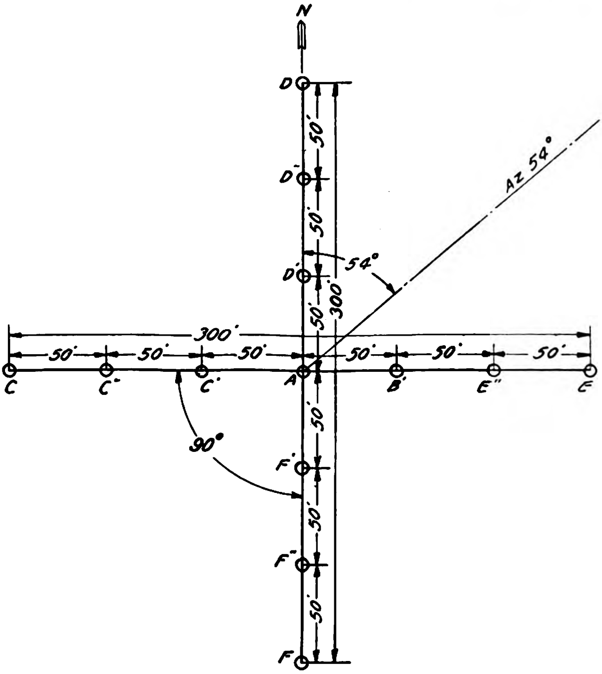

Fig. 6—Radio-Beacon Loop System, Hamakuapoko, Maui, T.H Lat. 20 deg. 55 min. N., Long. 156 deg. 20 min. W.

The following day orders were issued for me to proceed to Kahului, Maui, via steamer for temporary duty in connection with the selection of a site for the radio-beacon station. The small inter-island steamer left Honolulu Sunday evening and arrived at Lahaina on the south side of Maui at 4 o'clock on the following morning. Getting ashore was accomplished by means of one of the life boats, as docking facilities were not available. A 20-mile automobile trip was then necessary to reach the town of Wailuku, where hotel accommodations were available. Here a local map was obtained and a "Drive Yourself" car hired for the day. The road along the shore from Kahului to Huelo was followed and a careful search made for an open and fairly level location. The terrain in this direction was found to be very rugged and, therefore, unsuitable, except on the edge of the cliff near Maliko. This location, however, was undesirable, as it was too far from Kahului where the equipment would be unloaded. A reconnaissance was then made inland between Haiku and Paia, where there are many sugar cane and pineapple fields, and towards noon a very desirable field was located at Hamakuapoko. Permission to establish the radio-beacon station temporarily was obtained from the owners. Upon returning to the hotel that evening I was fortunate in meeting a territorial civil engineer who was familiar with the location of triangulation stations in the vicinity of Hamakuopoko. The next day this engineer was engaged to survey and lay out the loop system.

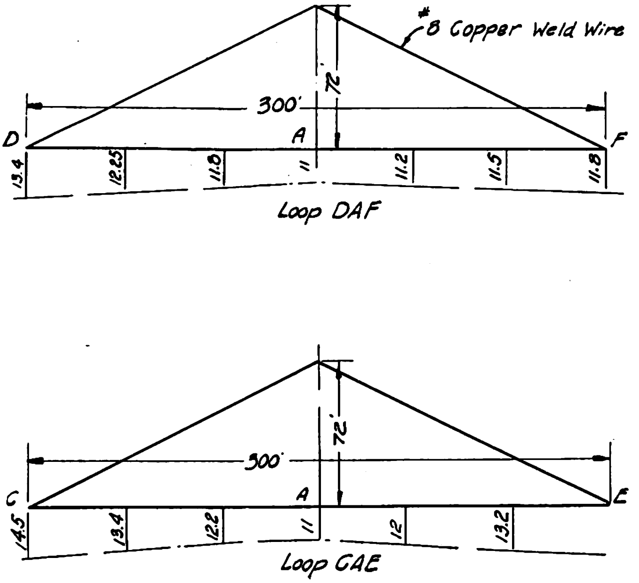

Fig. 7—Radio-Beacon Loop System, Hamakuapoko, T. H.

The position of the loops was carefully laid out and marked with stakes, as shown in Figs. 6 and 7.

Upon arrival at Fort Shafter the following morning a radiogram was received advising that the Crissy Field beacon would be operated for test from 11:00 P.M. to 1:30 A.M. (P. T.) or 8:30 P.M. to 11:30 P.M. (Hu. T.). We listened during this period but could not hear the Crissy Field beacon signals at Fort Shafter. Naturally we were disappointed with the results, but not altogether discouraged, as receiving conditions in that locality were considered very poor. The remainder of the week was spent in gathering additional supplies and arranging for enlisted personnel to assist in the construction of the beacon station on the Island of Maui. A pole suitable for supporting the loops was obtained from the Navy Department at Pearl Harbor and towed to the inter-island wharf at Honolulu.

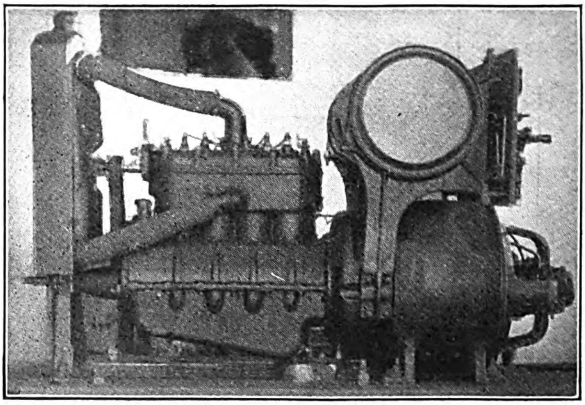

Fig. 8—Power Equipment Type PE-40. Gasoline-Engine-Driven 5 kw Direct-current Generator.

On June 7, 1927, accompanied by a detachment of four enlisted men from Luke Field, the writer left Honolulu for Kahului on the SS *Mauna Kea*. All of the equipment and supplies were on board, including the 93-foot pole which was lashed to the rail. We arrived on the following morning and quarters were obtained on the plantation of the Maui Agricultural Company. Work on the construction of the loops and of a building was started immediately.

On the following day a temporary radio station was established for communication with the Department Signal Officer at Fort Shafter, Honolulu. An airplane radio transmitter type BC-114[4] (SCR-134 set) was used for transmitting and a receiver type SE-1420, with two stages of audio amplification, was used for receiving. The radio station at Fort Shafter (FX-2) transmitted on a wavelength of 1600 meters. The Maui station (EH-9) transmitted on a wavelength of 450 meters, or 750 meters, depending upon receiving conditions at Fort Shafter. The distance between Fort Shafter and Hamakuopoko, Maui, is approximately 90 miles, and reliable communication was maintained on regular schedules daily.

During the week of June 7 the Crissy Field beacon carried on regular tests with the SS *Manoa* of the Matson line, sailing between San Francisco and Honolulu, which were very successful. The Crissy Field beacon was heard at Hamakuopoko, Maui, on an SE-1420 receiver using two stages of amplification, with a signal strength of R-4 to R-5 during the hours of darkness. The letter *N* was quite pronounced, superimposed on the background character, indicating that the bearing of the equisignal zone being transmitted from Crissy Field was slightly to the south of the correct course. It had been arranged previously that each station would be so adjusted that the *N* signals would be heard in an airplane when it was north of the correct course and the *A* signals when south of the course. This was the first time that the Crissy Field beacon had been heard in the Hawaiian Islands, and the results were most gratifying, as they not only showed that the modified transmitter had a much greater range than the old-type spark sets, but also made it pos-

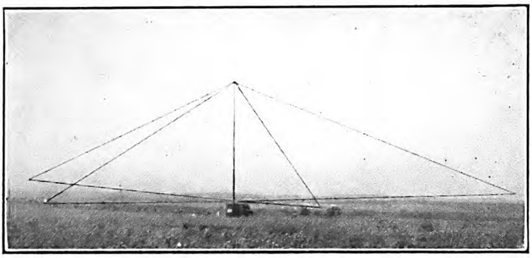

Fig. 9—Exterior of Radio Beacon, Maui, T. H. Mt. Haleakala in Background.

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[4] This transmitter is designed for telephone, C. W. and tone telegraph. It consists of three 50-watt tubes: master oscillator power amplifier and modulator; also, one 5-watt speech amplifier. Range: telephone—30 miles; C. W. telegraph—100 miles.

sible to adjust readily the transmitted goniometer bearing upon the advice of the listener at the opposite end of the course. In other words, the goniometer bearing could be shifted so as to establish a bracket on the receiving station at the opposite end of the course in order that the average of these readings could be used in the final setting of the goniometer.

The Crissy Field beacon was again operated on the night of June 14th from 10:30 P.M. to 11:40 P.M. (P. T.). The signals were barely audible from 8:30 to 8:40 P.M. and then disappeared entirely during an eclipse of the moon. On the following night this station was again operated on the same schedule with the goniometer set on 240 deg. The results of this test indicated that the bearing was still slightly south of the true course and that the keying was rather slow with poor interlocking. The signals occurred at about seven-second intervals with a signal strength of R-4.

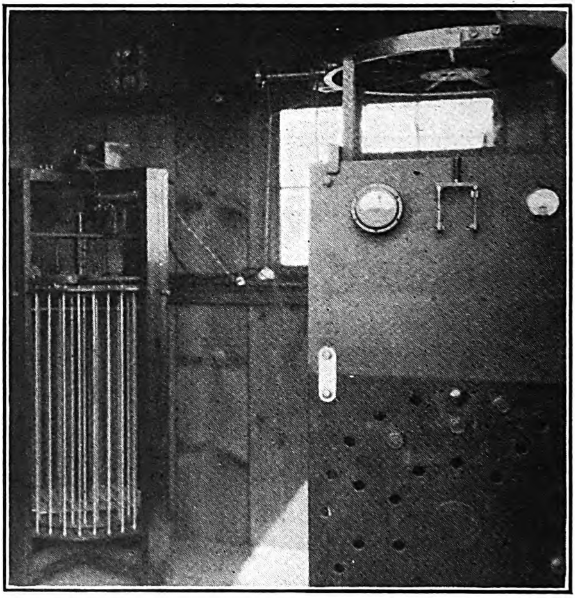

Fig. 10—Interior of Radio-Beacon Station at Hamakuapoko, Maui, T. H

Further tests were carried on, and on the night of June 21 the bearing was increased to 241 deg.[5] During this test, and a similar one on the following night, a broken dash, which might be described as a letter *X*, was received, which showed the bearing of 241 deg. to be correct. The interlocking, however, was slightly defective, probably due to adjustment of the automatic key or relay. At this time a standard superheterodyne receiver type BC-115–116 was also used for comparison with the SE-1420 receiver. Prior to the beacon tests broadcasting stations on the Coast were heard with very good volume. However, this set was much less sensitive on high wavelengths, (1035 meters) and the beacon signals could not be heard with it, although they came in R-4 to R-5 with the SE-1420 receiver, with regeneration just below the point of oscillation. The operator at Crissy Field was then advised that the bearing was correct from Crissy Field to Hamakuopoko, Maui. It was also advised that the BC-115–116 receiver was not suitable for long flights with the radio beacon.

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[5] Considerable distortion caused by a stray coupling within the goniometer was observed at the Crissy Field Station. This accounts for the fact that the bearing as indicated on the goniometer scale did not coincide with the computed bearing.

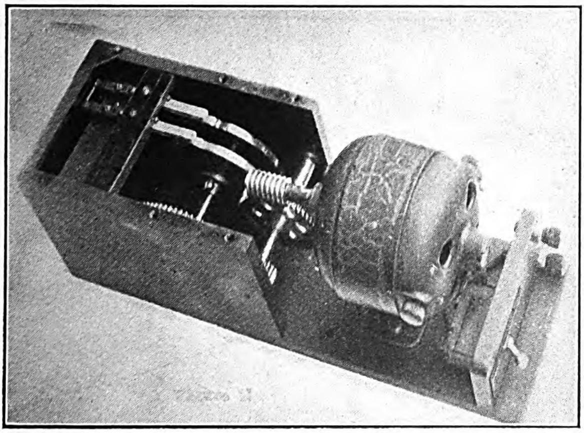

Fig. 11—Automatic Keys for Radio Beacon.

The installation of the Maui radio beacon was completed on June 21st and preliminary test showed it to be well-balanced and ready for operation. Fig. 9 shows the exterior of this station with Mt. Haleakala in the background. Fig. 10 shows the interior with goniometer. The goniometer was set on the computed bearing of 52 deg. 31 min. for the preliminary tests. Both stations were tested on the night of June 23rd. The observer at the San Francisco station reported the Maui beacon signals were heard with signal strength R-3 on an SE-1420 receiver through static and some interference. The letter *A* predominated during this test, which indicated that the bearing being transmitted by the

before leaving McCook Field, Dayton, Ohio enroute to San Francisco via Chicago, Fort Sill, San Antonio, El Paso and San Diego. The Crissy Field beacon was used during the flight from San Diego to San Francisco in order that the pilot and

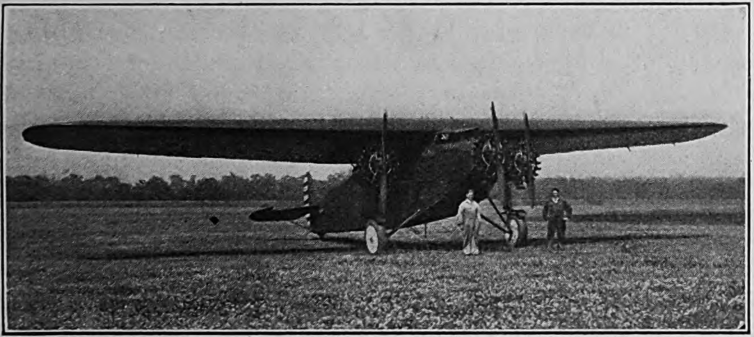

Fig. 13—Fokker Monoplane Type C-2, Used by Lieutenant Hegenberger and Lieutenant Maitland in Flight from San Francisco to Hawaii, June 28 and 29, 1927.

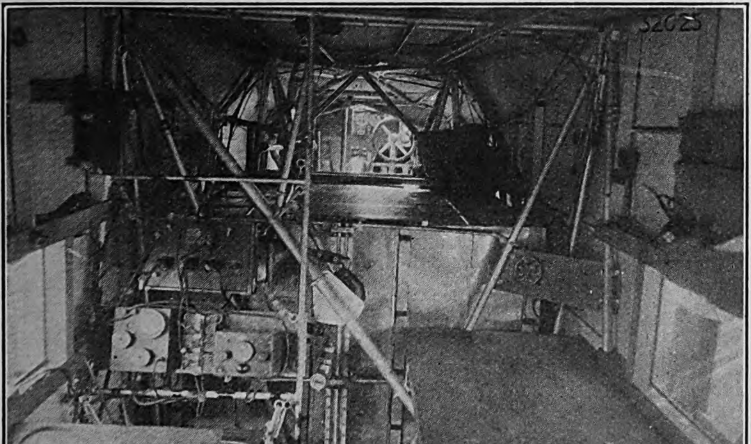

Fig. 14—Interior of Airplane Type C-2 Used by Lieutenant Hegenberger and Lieutenant Maitland in Flight from San Francisco to Hawaii. The SCR-134 Radio Telephone and Telegraph Set Is Shown Mounted in Special Framework on Left Side of the Cabin. The Table on the Right Was Used As A Chart Table for the Navigator.

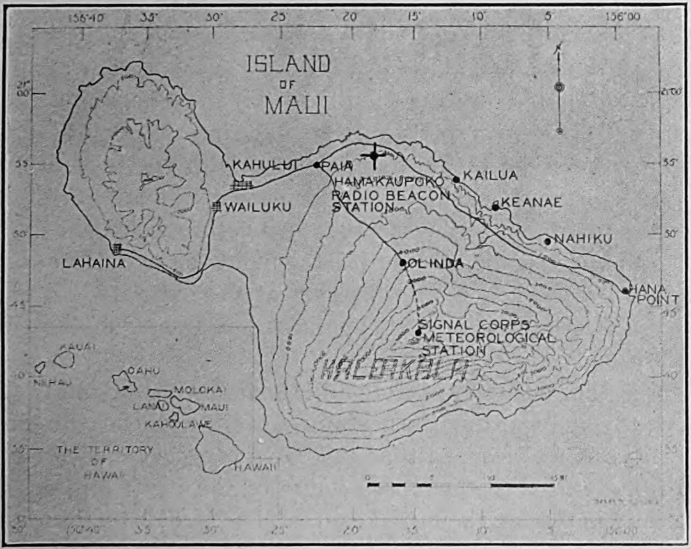

Fig. 16—Chart of Island of Maui Showing Location of Radio Beacon Station and Meteorological Station on Mt. Haleakala.

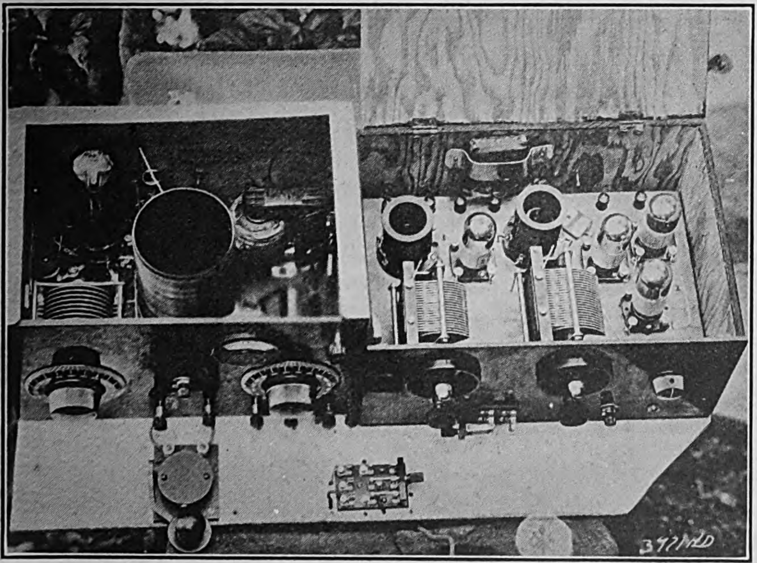

Fig. 17—Fort Shafter, T. H. August 19, 1927. Radio Set Used by Goebel in the Plane *Woolaroc*, Which Flew from San Francisco to Wheeler Field, T. H.

ports on the progress of the airplane were sent out by ships so that the position of the airplane was known throughout the flight.

The Maui beacon station was placed in operation at 5 P.M. on June 28, as it was estimated that the airplane would be approximately half way to Hawaii at that time. Starting at midnight weather reports were transmitted with the beacon every hour by substituting a hand key in place of the automatic key. The station functioned perfectly during the entire run of approximately fifteen hours. During the flight the signals from each station were checked by an observer at the opposite end of the course and no appreciable change in the character of the transmitted signals was noticed.

The author returned to Honolulu the following day and talked with the flyers regarding the usefulness of the radio beacon during the flight. It was learned that the San Francisco beacon signals were followed for the first two hours of the flight and assisted materially in setting the initial course. The signals then suddenly ceased (this was due to failure of gasoline engine at the Crissy Field beacon station) and were then picked up intermittently during the remainder of the flight. When about midway between San Francisco and Hawaii the Maui beacon signals were heard by Lieutenant Hegenberger who estimated the width of the *T* signal zone to be approximately eight miles. The beacon signals, however, were not heard continuously and it was believed that something was defective with the receiving equipment on board the airplane. A test flight made subsequent to the airplane's arrival at Wheeler Field, however, did not reveal any defects in the receiving equipment. It is regrettable that an expert radio operator could not have been included in the crew so that the radio equipment and reception of the signals could have received the entire attention of one man.

Part B

THE SMITH-BRONTE FLIGHT

July 14–15, 1927

The next flight in which the Army radio beacons were used was that made by Mr. Ernest Smith, pilot, and Mr. Emory Bronte, navigator in a single-motored "Travel Air" monoplane. The radio equipment installed in this airplane was made up hastily at San Francisco and is described below:

Transmitter. Hartley Circuit using one 50-watt tube and tuned to 600 meters for use in communication with ship stations.

Plate and filament supplied from a 900-cycle wind-driven generator, type GN-4.

Weight of transmitter complete approximately 50 pounds.

Receiver. One-stage tuned radio-frequency and regenerative detector with two stages of audio amplification. Plug-in coils to cover wave band from 500 to 1100 meters. Regeneration controlled by detector filament rheostat. Type DV-3 tubes, with dry cells for filament supply were provided. Weight of receiver approximately 15 pounds.

The above equipment was installed in the fuselage in back of the navigator's cockpit. The ignition system of the single Wright Whirlwind engine was not shielded, but the receiver was far enough away from the engine so that comparatively little ignition interference was present. The call letters of this airplane were WLO.

On July 14, 1927, the take-off was made from Oakland Airport at 10:40 A.M. with Honolulu as the destination. The log of the Crissy Field radio-beacon station for this flight is shown below:

July 14	
10:30 P.M.	Started beacon.
11:10 A.M.	Smith's plane passed over beacon station.
1:00 P.M.	Airplane reports beacon signals received O.K.
1:10 P.M.	Airplane reports still foggy—Beacon coming in fine.
2:00 P.M.	Still hearing beacon signals.
3:25 P.M.	Bronte reports signals not heard since 2:24 P.M. Thinks receiver out of order.
July 15	Bulletins came in during night but no report of beacon signals.
3:00 A.M.	Beacon shut down.
	Signals from Maui received at Crissy Field about R-2.
7:45 A.M.	Plane reported landing in sea 600 miles from Honolulu. Out of gas.
12:16 P.M.	Plane reported landing on Molokai—wrecked.

The Maui beacon station started transmitting when the airplane was estimated to be about half-way between San Francisco and Hawaii. In accordance with previous arrangements this station suspended its regular transmission of bearing for the first ten minutes of each hour to enable the operator to listen for the signals from the Crissy Field beacon. At every hour, on the half hour, the Maui beacon again suspended its regular beacon transmission while meteorological data was transmitted on one of the loops.

The log of the Maui radio-beacon station is shown below:

July 14.		
3:45 P.M.	(H.T.)	Started beacon 4.5 kw input.
3:45 P.M.	"	Message relayed from plane—Lat. 36.32 N. Long. 123.33 W. Still foggy have not heard beacon for last hour.
8:10 P.M.	"	Position of plane at 6:00 P.M. (P.T.) Lat. 35.08 N. Long. 130.46 W. Going fine.
9:00 P.M.	"	Crissy Field beacon heard R-4 heavy QRM.
11:00 P.M.	"	Crissy Field beacon heard R-4 heavy QRM.
11:55 P.M.	"	Crissy Field beacon heard R-4 heavy QRM.
July 15.		
12:15 A.M.	"	Received from Presidio S. F. "Maui beacon received here very good."
1:30 A.M.	"	Crissy Field beacon not heard. U.S.A.T. *Kenowis* reported through KHK (Honolulu) that he heard plane, WLO, send out a QST but was unable to copy due to QRM. (12:30 A.M. (Hu. T.)) (*Kenowis* about 700 miles from Honolulu).
2:00 A.M.	"	SS *Wilhelmina* (WMO) reported to Honolulu (KH) at 1:57 A.M. (H.T.) that plane's position report at 4:05 A.M. (P.T.) was Lat. 28.58 N. Long. 144.36 W.
2:15 A.M.	"	From Presidio WVY "Think Crissy Field beacon station closed now as unable to reach them by phone."
3:00 A.M.	"	Intercept direct from plane. Long. 145.45 W. All well aboard pretty tired too.
3:10 A.M.	"	Received relayed message from plane "We are out of gas."
3:24 A.M.	"	Intercept "Can you make it to Honolulu" WYAD working WLO.
4:00 A.M.	"	Received QST direct from plane (WLO) "Rush help to WLO making forced landing in water NE of Paia, Maui."
5:06 A.M.	"	Received QST direct from Plane (WLO) "500 miles north east Paia all out of gas, send help, planes, boats."
5:15 A.M.	"	Received direct from plane (WLO) "Landing in sea, have rubber life boat but send help."
5:49 A.M.	"	Received QST from plane (WLO) "Please send help to WLO coming down no gas."

6:20 A.M.	(H.T.)	Sent QST on beacon "Smith plane (WLO) probably been using radio beacons from Coast to Hawaii stop Letter *T* indicates course stop Letter *N* indicates north of course stop Letter *A* indicates south of course stop Wavelength ten thirty five meters.
7:25 A.M.	"	S.S. *Wilhelmina* (WMO) reported position about 75 miles from plane at 6:35 A.M. and proceeding to plane's position.
7:35 A.M.	"	Repeated QST sent out at 6:20 to assist rescue ship in proceeding to disabled plane.
7:55 A.M.	"	QST "Will resume transmitting beacon signals on San Francisco to Maui bearing at 9:00 A.M."
8:00 A.M.	"	Sent message to Signal Officer Ft. Shafter "Suggest following be broadcast on 600 meters 'Smith airplane was following radio beacon signals on ten thirty five meters from Coast enroute Honolulu stop Letter *T* indicates course stop letter *N* indicates north of course stop Letter *A* indicates south of course.'"
9:15 A.M.	"	At approximately 9:15 A.M. Smith's Plane was sighted over coast of Maui and headed towards Paia. When nearly over the receiving station at Hamakuopoko he turned towards Molokai and disappeared in the clouds. At about 11:00 A.M. a radiogram broadcast by a small privately-owned radio station on the island of Molokai advised that the airplane *City of Oakland* had landed in Kiawi trees on the southeastern shore of Molokai out of gasoline. Smith and Bronte unhurt.

A flight of Army airplanes from Luke Field, Oahu landed at an emergency field on the island of Molokai and took Messrs. Smith and Bronte to Honolulu. The wrecked airplane *City of Oakland* (Fig. 15) was salvaged by naval personnel and taken to the Naval Air Station at Pearl Harbor. A copy of a letter written by Mr. Bronte commenting upon the use of the radio beacon is shown below.

San Francisco, August 8, 1927.

Mr. P. Hendricks,
Crissy Field,
Presidio, San Francisco.

Dear Sir:

It gives me great pleasure at this time to thank you for the splendid cooperation given Mr. Ernest Smith and myself on our recent flight from the Oakland Air Port to the Hawaiian island in the monoplane *City of Oakland*. It is to be regretted that our headphones went out of commission when we were approximately

200 miles at sea because up to that time the radio beacon had functioned perfectly and we were desirous of using it on the entire flight.

I might say at this time that it was the San Francisco radio beacon which set us on our course from the coast, for, as you know, we left on the morning of July 14th in a dense fog and after climbing through the fog, found ourselves quite a bit to the northward over Marin County. At that time the beacon signal *N* was coming in very strongly and we headed due south until the signal *T* was heard most distinctly. We then set the plane on our course for the Hawaiian groups keeping the *T* signal tuned in meanwhile.

Personally I believe the radio beacon to be the greatest aid to long distance aerial navigation that is known today.

Thanking you again for your kind cooperation before and during the flight, I am

Sincerely yours,
(signed) E. B. BRONTE

EBB/RC

Part C

THE DOLE FLIGHT

August 16–17, 1927

The Honolulu Chapter of the National Aeronautic Association requested the War Department to authorize the use of the Army radio-beacon station at Crissy Field, San Francisco, and at Hamakuopoko, Maui, T. H., during the Dole Flight which was scheduled to start from San Francisco at noon August 12, 1927 with Wheeler Field, Honolulu, as the destination. As soon as the Hawaiian Department was notified that the above request had been granted preparations were made whereby the greatest service possible would be rendered by the radio-beacon system.

During the Army flight by Lieutenant Maitland and Lieutenant Hegenberger, as well as the Smith-Bronte flight from San Francisco to Hawaii, the radio beacons were used as an aid to aerial navigation by sending out signals to indicate the Great Circle course from Crissy Field to Hamakuopoko, Maui. This service was interrupted at intervals to transmit meteorological data which was obtained locally. However, due to trouble with the receiving apparatus in the airplanes on these flights, the aviators were not able to utilize to the fullest degree the assistance which was offered.

The Signal Officer, Hawaiian Department, realizing the possibilities of the radio beacon, not only as a guide along the correct course, but as a means of furnishing the aviators with reliable meteorological data, ordered a meteorological detachment to the summit of Mt. Haleakala on the island of Maui, shown in Fig. 16. The information obtained for several days prior to the Dole Flight was forwarded to the Weather Bureau at Honolulu and at San Francisco to assist in forecasting the weather during the flight. Telephone communication was established between the meteorological station on Haleakala and the radio-beacon station at Hamakuopoko.

Owing to the fact that only a very few of the entrants for the flight were ready for the start at noon of August 12, 1927, the start of the Dole flight was postponed until noon of August 16. The lineup of the planes entered on that date was as follows:

Plane No.	Call Letters	Frequency of Transmitter Kilocycles	Power Watts	Beacon Receiver	Type Plane	Name of Plane	Pilot	Navigator
1.	KDE	500		Yes	Monoplane	*Oklahoma*	Griffin	Henley
2.	KNK	500	100	No	Monoplane	*El Canto*	Goddard	Lt. K. C. Hawkins, U.S.N.
3.	KGGA	9050	50	No	Monoplane	*Pabco*	Maj. Irving	None
4.	KW5	None		Yes	Monoplane	*Golden Eagle*	Frost	Scott
5.	None	None		No	Biplane	*Miss Doran*	Pedlar	Knope
6.	None	None		No	Monoplane	*City of Peoria*	Parkhurst	Lowes
7.	KGGI	493	50	Yes	Monoplane	*Woolaroc*	Goebel	Lt. Wm. V. Davis, U.S.N.
8.	None	None		No	Monoplane	*Spirit of Dallas*	Erwin	Eichwaldt
9.	None	None		No	Monoplane	*Aloha*	Jensen	Schlueter

The first airplane took off at Oakland Airport at 12:30 P.M. (P. T.). Several of the planes had trouble after getting into the air, forcing them to return, while others were unsuccessful in getting off the ground. Only four of the nine airplanes entered for the race actually started out. They were numbers 4, 5, 7, and 9. Two of these, No. 4 and No. 7 were equipped with beacon receivers and No. 7 also carried a transmitting set. The radio equipment installed in the *Golden Eagle*, entrant No. 4, was essentially the same as that of the *Woolaroc* which will be described below. The transmitter, however, was removed before the flight to decrease the weight of the airplane.

The radio equipment installed in the *Woolaroc,* entrant No. 7, consisted of:

> *Receiver.* One-stage tuned radio frequency, regenerative detector and two stages of audio amplification. This was the same receiver that was previously used in the Smith-Bronte flight to Honolulu.
>
> *Transmitter.* Same one used in Smith-Bronte flight to Honolulu. 600-meter ACW Hartley Circuit using 50-watt tube. Filament and plate supplied from 900-cycle wind-driven generator, type GN-4.
>
> Note: A dual-antenna installation was made so that if one was lost another could be used. Fig. 17 shows transmitting and receiving sets and Fig. 18 shows the schematic wiring of the complete installation.

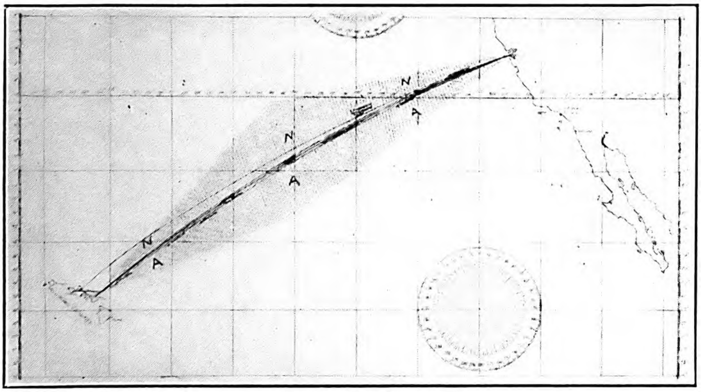

Fig. 19—Radio Beacon Chart—Dole Flight. Chart Showing Great Circle Course Transmitted by Radio Beacons between San Francisco and Maui, T. H. The Width of the Equi-Signal Zone at Approximately Half-Way between These Two Points Was Estimated to be about Eight Miles from Observations Made During Flight and on Surface Vessels.

The Crissy Field beacon was placed in operation at 12 o'clock noon of August 16th and continued in operation until 11:00 A.M. August 17th. No trouble of any kind was experienced with this set during the entire period of twenty-three hours.

The Maui beacon was placed in operation at 7:30 P.M. (Hu. T.) or approximately ten hours after the start of the flight, as it was assumed that the San Francisco beacon would operate during the first twelve hours of the flight which would make a two-hour overlap in transmission. This station functioned perfectly with full power during the entire night and starting at daybreak, meteorological data consisting of wind direction and velocity, cloud density, cloud ceiling, cloud summit, and direction and velocity of wind drift component was transmitted during the first ten minutes of each hour.

Position reports from the *Woolaroc*, which were received at frequent intervals, indicated that this airplane was following a direct course from San Francisco to Hamakuopoko, Maui, T. H., as shown in Fig. 19. It was also reported that signals from both of the beacon stations were heard and that the *T* signals were being followed continuously.

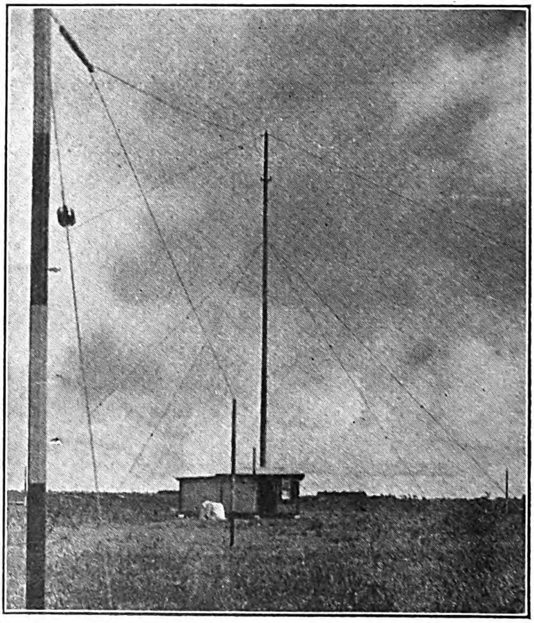

Fig. 20—Radio-Beacon Station at Wheeler Field, Oahu, T. H. The Airplane Hangars May Be Seen in the Background.

The *Woolaroc* landed at Wheeler Field at 12:24 P.M. (Honolulu time) thereby winning the first prize of $25,000.00. Messrs. Goebel and Davis were taken into Honolulu immediately, where each gave a short talk on their flight from a local broadcasting station. A copy of a letter written by Lieutenant Davis commenting upon the flight is shown below.

San Francisco, California,
1 Sept. 1927.

Signal Officer,
9th Corps Area Headquarters,
San Francisco, Calif.

Dear Sir:

I wish to express my appreciation of the Army courtesies to Art Goebel and myself in conjunction with the Dole Flight. The Army was of great assistance in many ways and I cannot tell you how we appreciated it. In particular, the radio beacon was an unqualified success, and rendered the navigational problem a very simple one. The signals were clear and distinct and were held practically the entire way. At one time I was able to hear both the San Francisco and Maui beacons at the same time, our position at the time being about midway between Honolulu and San Francisco.

I will write a detailed report of the flight upon my return to duty and will be glad to forward a copy to you, if it would be of interest.

Thanking you again, I remain

Respectfully yours,
(Signed) William V. Davis
Lieut. (j.g.) U.S.N.

Naval Air Station
San Diego, Cal.

On August 24 the author returned to Honolulu and had the opportunity of talking with Mr. Goebel and Lieutenant Davis, pilot and navigator respectively of the *Woolaroc.* Both of these gentlemen were very enthusiastic in their praise of the radio beacons which had so greatly simplified their navigation problems. Lieutenant Davis said that the San Francisco beacon signals faded out for about two hours before sunset, but the Maui signals were picked up immediately when started at 7:30 P.M. (Hu. T.). After sunset the San Francisco beacon signals were again heard and for a period both beacon stations were heard simultaneously. The position of the *Woolaroc* at that time was about midway between San Francisco and Honolulu and the width of the *T* zone was estimated to be about 8 miles. The islands were sighted when about 40 miles from the coast of Maui and the beacon course was then abandoned and a new course was set towards Wheeler Field on the island of Oahu. Towards the latter part of the flight a connection in one of the receiver tube sockets became crystallized and broke off. Lieutenant Davis located the trouble, made a temporary repair connection and then held the tube in place for about two hours, when the islands were sighted and the course changed. Lieutenant Davis also said that in obtaining his position at various times he calculated the distance from San Francisco along the beacon path by dead reckoning and then obtained the latitude by sighting on Polaris. This latitude line when projected to intersect the Great Circle beacon course gave the longitude of the airplane at that time.

The second prize of $10,000.00 was won by Martin Jensen, pilot of the *Aloha*. The two remaining airplanes which started in the Dole Derby were lost, and nothing has since been heard of either of them. The *Spirit of Dallas* with Wm. Erwin, pilot, and Eichwaldt, navigator, left San Francisco enroute to Honolulu on August 18th. This airplane carried a short-wave transmitter. The last report heard of this airplane was when it was in a tail spin about six hundred miles from San Francisco.

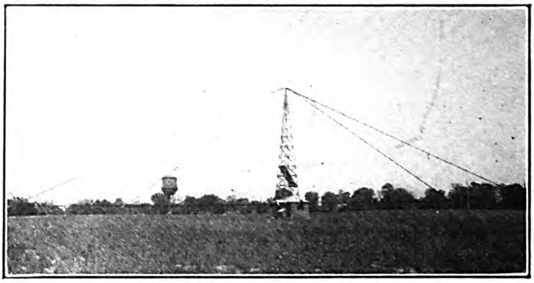

Fig. 21—Radio Beacon, Wright Field. Building, Tower, and Loop System.

In the final analysis of the data which was obtained during these tests the following conclusions are set forth.

(1) That the equi-signal radio beacon is already developed to the point where it is capable of furnishing important assistance in aerial navigation.

(2) That reception aboard the airplane is the weakest part of the beacon system.

(3) That with the best transmitting and receiving apparatus adequate personnel training is necessary, as it was demonstrated in these flights that in one case excellent training of the radio operator made up for deficiencies in apparatus while in another case with adequate apparatus a comparatively untrained operator utilized the beacon only part of the time.

when about 1100 nautical miles from Hawaii, which checks rather closely with the estimates made by Hegenberger and Davis on their respective flights.

IMPROVEMENTS IN THE RADIO BEACON

During the past year several improvements have been made in the radio beacon by the Signal Corps Aircraft Radio Laboratory at Dayton, Ohio. It has been observed that the average pilot has more or less difficulty in differentiating between the *N* and *A* signals when the airplane gets slightly off the course. To overcome this difficulty, the letters *N* and *A* have been superseded by groups of dots and dashes. These signals interlock as before and the pilot hears a continuous 1000-cycle note when on the course. A series of dots or dashes is heard when the airplane gets off the correct course. This system has been found to be superior in that the zone of interlocked signals is made narrower and to a more rapid change in the quality of the signals.

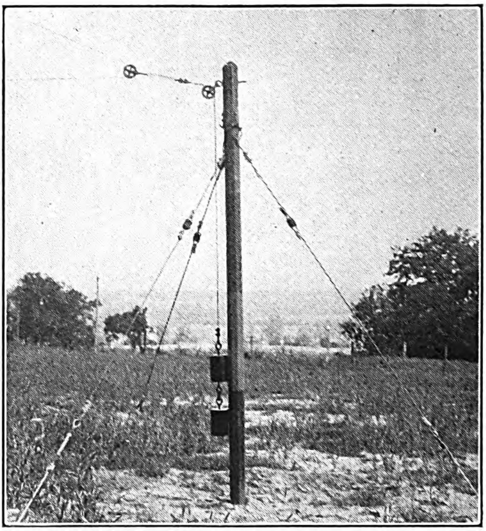

Fig. 23—Radio Beacon, Wright Field. Detail of End of Loop.

The automatic key and switching relay in the rotor circuit of the goniometer has also been improved so that both are much more simple to adjust and are rugged enough to maintain their adjustments. A complete new installation has been made at Wright Field as shown in Figs. 21 and 22. Fig. 23 shows methods of holding loops in position. A master oscillator and neutralized power-amplifier circuit is used to obtain a more constant frequency, as it was observed that there was a tendency for the frequency to shift when the goniometer was coupled directly to the oscillator. In the new installation the transmitter is entirely shielded to prevent any stray coupling between it and the goniometer or loop circuits. The results obtained during test flights have been very satisfactory. Recently flights have been made from Dayton to Buffalo, Cleveland, Detroit, Uniontown, and St. Louis. During these flights the equi-signal zone appeared to be approximately 1 deg. in width.

Volume 16, Number 9 *September, 1928*

EFFECTIVE HEIGHTS OF THE KENNELLY-HEAVISIDE LAYER IN DECEMBER, 1927 AND JANUARY, 1928*

By

G. Breit, M. A. Tuve, and O. Dahl

(Department of Terrestrial Magnetism, Carnegie Institution of Washington, Washington, D. C.)

Summary—*Effective heights of the Kennelly-Heaviside layer have been measured from December 19, 1927 to January 16, 1928. It is found, with the improved conditions, that the heights of successive reflections are approximately in the ratios 1, 2, 4.*

THIS is a brief report on the results obtained by the echo-method. The transmission used for this has been improved so as to obtain sharp peaks. The details of the method have been described by M. A. Tuve and O. Dahl.[1] The sharpness of the peaks obtainable and the large spacing between them make the identification of the ground-wave and the reflected wave much more certain. The accuracy with which the time of arrival of the two waves can be measured is increased as well. There are two reasons for this: (1) The graphical accuracy of locating the peaks is higher, and (2) the interference between the successive waves is completely eliminated and hence no correction for relative phases of interfering waves has to be made.

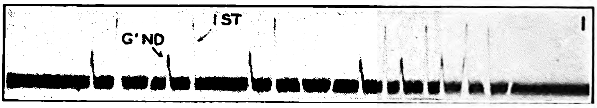

Fig. 1—Specimen Test Records; Test 88 during $9^h 50^m$ to $10^h 15^m$, December 29, 1927.

As in our previous experiments the transmission took place from the Naval Research Laboratory located at Bellevue, Anacostia, D. C. The signals were received at the laboratory of the Department of Terrestrial Magnetism about 7 miles northwest from the transmitter. The frequency used was 4015 kilocycles.

Some representative wave-forms are shown in Figs. 1, 2, 3, and 4, for all of which the time-scale is increasing from left to right. In Fig. 1 a case of single reflection is shown. The approximate effective height is 140 miles. The smaller hump is due to the ground, the larger to the reflection. In Fig. 2 a case of double reflection is shown. The strong reflection gives a height of 140 miles; the weak one corresponds to 280 miles. These wave-forms represent the usual conditions during the period discussed.

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* Original Manuscript Received by the Institute, May 23, 1928. Communication from the International Union of Scientific Radiotelegraphy.

[1] Proc. I. R. E., 16, p. 794; June, 1928.

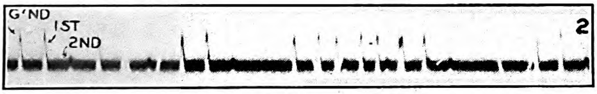

Fig. 2—Specimen Test Records; Test 88 during $9^h\ 50^m$ to $10^h\ 15^m$, December 29, 1927.

Under exceptional conditions triple reflections are observed. Examples are shown in Figs. 3 and 4. The wave-forms were obtained on January 14 during a period of severe short-distance fading. The three reflections correspond to heights of 67, 140, and 280 miles. As in most cases the amplitudes of the reflections are different in the two cases while the strength of the ground-wave remained constant.

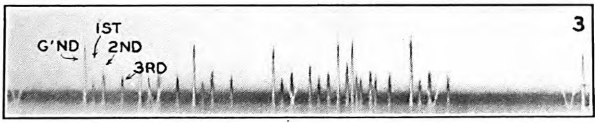

Fig. 3—Specimen Test Records; Test 109 during $9^h\ 30^m$ to $9^h\ 45^m$, January 14, 1928.

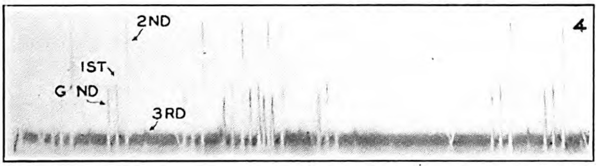

Fig. 4—Specimen Test Records; Test 109 during $9^h\ 30^m$ to $9^h\ 45^m$, January 14, 1928.

We have not observed such multiplicities in our previous tests. We had cases of multiple reflections, but the retardations were not different from each other by factors of two as they are here. Heising[2] observed retardations having integral ratios for considerably larger total time-differences. This is presumably due to the fact that our older tests did not resolve the reflections

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[2] Proc. I. R. E., **16**, p. 75, January, 1928.

TABLE *I*

EFFECTIVE HEIGHTS OF LAYER AND RELATIVE INTENSITIES OF REFLECTIONS AT WASHINGTON, DECEMBER 1927 TO JANUARY 1928

Test No.	Date	75° west M M.T.		Heights			Intensities			
							Ground	Reflection		
		From	To	First	Second	Third		First	Second	Third
	1927	h m	h m	miles	miles	miles				
74	Dec. 19	9 30	10 00	135	285	—	—	—	—	—
				135	270	—	—	—	—	—
				135	—	—	—	—	—	—
75	Dec. 19	14 25	15 05	140	—	—	—	—	—	—
				140	—	—	—	—	—	—
				145	—	—	—	—	—	—
				145	—	—	—	—	—	—
76	Dec. 20	9 30	10 00	145[a]	285[a]	—	4	5	0.5	—
				135[a]	—	—	3.5	4	0	—
				135[b]	—	—	7	1.5	—	—
				135[b]	—	—	7	1.5	—	—
78	Dec. 21	9 30	9 55	140	—	—	5	4	—	—
				140	—	—	3	1.5	—	—
				140	—	—	3	2	—	—
79	Dec. 21	14 30	15 00	140[c]	—	—	—	—	—	—
80	Dec. 22	9 30	10 00	134	272	—	—	—	—	—
81, 82	Dec. 22	14 30	15 00	46?	93?	140	5	—	—	2
86, 87	Dec. 28	9 30	9 55	75	140	—	5	3	2.5	—
				140	—	—	4	4	—	—
		14 30	14 50	140	—	—	3.5	1.4	—	—
				140	—	—	5.0	3.0	—	—
				140	—	—	4.4	4.3	—	—
88	Dec. 29	9 50	10 15	140	280	—	3.2	6	0.5	—
				135	—	—	3.5	8	0	—
				140	280	—	3.5	4.0	0.5	—
				135	—	—	3.2	3.2	—	—
				135	—	—	3.9	1.0	—	—
89, 90	Dec. 29	14 30	14 45	145	—	—	3.2	8.5	—	—
91	Dec. 30	14 30	15 00	145	—	—	6.0	0.8	—	—
				140	—	—	7.5	0.5	—	—
	1928									
92	Jan. 3	14 30	14 50	142	—	—	7	3	—	—
93	Jan. 4	9 30	9 50	132	—	—	3.5	2.8	—	—
				132	—	—	4.3	5.0	—	—
	1928	h m	h m	miles	miles	miles				
94	Jan. 4	14 30	14 50	140	—	—	5	3	—	—
95	Jan. 5	9 30	9 45	145	—	—	6.5	12.5	—	—
				140	285	—	6.5	10.6	1.0	—
97	Jan. 6	9 30	9 50	140	—	—	2.0	5.5	—	—
98	Jan. 7	9 30	9 50	135	—	—	5	11	—	—
				140	280	—	5	11	1	—
100	Jan. 9	9 30	9 45	140	280	—	9	8	0.5	—
101	Jan. 9	14 30	14 50	145	—	—	9	3.5	—	—
102	Jan. 10	9 30	9 50	140	280	—	10	9	1	—
				140	280	—	10	15	1	—
				140	280	—	9	3	2	—
103	Jan. 11	9 30	9 50	140	—	—	10	12	—	—
104	Jan. 11	14 30	14 50	140	—	—	7	11	—	—
105	Jan. 12	9 30	9 45	75	150	—	8	10	14	—
				75	150	300	8	10	14	3
				75	150	—	8	6	13	—
106	Jan. 12	14 30	14 45	140	300	—	5	3	2	—
				140	300	—	5	5	0.5	—
107	Jan. 13	9 30	9 40	140	290	—	9	9	2	—
				140	—	—	9	13	0	—
				140	—	—	9	23	—	—
				140	280	—	9	3	2	—
				140	280	—	9	21	2	—
108	Jan. 13	14 30	14 45	140	280	—	3	8	1	—
				140	—	—	3	13	0	—
				140	—	—	3	5	0	—
				140	—	—	3	7	0	—
109	Jan. 14	9 30	9 45	65	140	280	9	12	11	1
				65	140	280	9	7	1	1
				70	140	280	9	2	3.5	3
				70	140	280	9	11	18	2
				70	140	280	9	9	19	1
110	Jan. 16	9 30	9 45	140	280	—	6	19	2	—
				140	280	—	6	2	1	—

[a] Loop. [b] Antenna. [c] Questionable 93-mile reflection.

completely. The interference between overlapping waves must have produced misleading effects.

Table I gives the measured effective heights and also rough estimates of intensities of the waves recorded.

The authors should like to express again their gratitude to Messrs. A. H. Taylor, M. H. Schrenk, and L. A. Gebhard for the possibility of using the *NKF* transmitter in these tests.

Volume, Number 9 *September, 1928*

CONSIDERATIONS AFFECTING THE LICENSING OF HIGH-FREQUENCY STATIONS*

By

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ADVENT of the use of high frequency in commercial fields puts us face to face with new problems of regulation, not entirely paralleled in previous radio situations, due to (1) the great economic value of these frequencies because of the comparatively low cost for long distance communication, and (2) because of the fact that a large share of these frequencies require, generally speaking, a world-wide clear channel. The following considerations appear to me pertinent in the allocation of frequencies between 6,000 and 23,000 kilocycles.

(1) Generally speaking, these frequencies must be considered as interfering around the world. Attention is called to the recognition of the long-distance value of these frequencies in the International Radio Convention Regulations with recommendation that they be reserved for long-distance communication in services between fixed points.

For this reason, frequencies assigned by one nation must not be later assigned by another nation unless it be demonstrated that, due to geographical separation and difference in day and night conditions, this is practicable without interference.

The distances which high frequencies carry may be summed up (roughly) in Table I:[1]

TABLE I

		Average Distance	
		Day-Miles	Night-Miles
1500–2000	kc	100	250
2000–2250	"	125	300
2250–2750	"	150	500
2750–2850	"	150	550
2850–3500	"	250	600
3500–4000	"	300	1000
4000–5500	"	450	2500
5500–5700	"	500	3500
5700–6000	"	550	4000
6000–6150	"	600	5000 +
6150–6675	"	800	"
6675–7300	"	1000	"
7300–8550	"	1200	"
8550–8950	"	1500	"
8950–9500	"	1800	"

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* Original Manuscript Received by the Institute, April 24, 1928. Revised Manuscript Received May 31, 1928.

† Technical Advisor to Federal Radio Commission.

[1] Author's Note: The distances are by no means fixed or constant, and only generally represent the situation, being dependent upon various factors such as different geographical locations.

9500–11000	“	2500	“
11000–11700	“	3500	“
11700–12300	“	4000	“
12300–12825	“	5000	“
12825–14000	“	5500	“
14000–23000	“	7000	Not useful at night

Frequencies above 6,000 kc have a silent zone due to the “skipping” of the wave, at distances varying up to hundreds of miles in the daytime and even thousands of miles at night, depending on the frequency from the transmitting station. This table is subject to deviation due to summer and winter, day and night, because of differences in the height of the Heaviside layer. Fig. 1 illustrates this.

The curves shown in Fig. 1 are plotted from the data received from the following sources: (a) American Telephone and Telegraph Company, (b) Radio Corporation of America, (c) General Electric Company, (d) U. S. Army, (d) Naval Research Laboratory, (f) actual operating data in the U. S. Navy. These curves

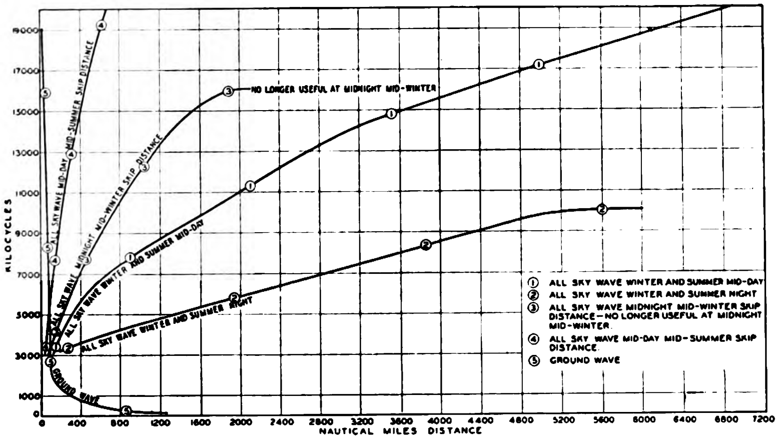

Fig. 1

are an approximation of the effective mid-day sky wave for all seasons, a night sky wave for all seasons, a skip distance curve for summer mid-day and winter midnight, and an all season ground wave for night and day. The curves are based on 1 kw in the antenna, CW telegraphy, using moderately sensitive receivers. Curve 1 shows the approximate distance versus frequency for an all season mid-day sky wave. Example: Given a distance 2200 miles find the frequency to produce a good readable signal at the receiving station. Vertically up from 2200 miles to Curve 1 and horizontally to the left read 11,400 kc. Given a certain frequency the distance will be found by the reverse process. The frequency is probably accurate to ±500 kc and the distance to ±200 miles, when in an east-west direction. The frequency and distance for north and south transmission may vary greatly from the above.

Curve 2 shows the approximate distance versus frequency for the night sky wave at all seasons. The curve is read in the same manner as explained above. To produce a good readable signal at a receiving station 2200 miles away when both the receiving and transmitting stations are in darkness, it is found that approximately 6000 kc will suffice. Also it is probable that any frequency between Curve 1 and Curve 2 at 2200 miles will suffice and the choice of frequency depends upon the experiment and the time of day. Greater distances are covered by the lower frequencies at night as shown by this curve. The frequency and distance for north and south transmission may vary somewhat from the above.

Curve 3 is an approximation of skip distance versus frequency for midnight mid-winter. The skip distance is measured horizontally to the right from Curve 5. The skip distance at any particular time of day or year may vary between the horizontal on Curves 3 and 4.

Curve 4 is an approximation of the skip distance versus frequency for mid-day mid-summer. The skip distance is measured horizontally to the right from Curve 5. Curves 3 and 4 are read in the same manner as described above.

Curve 5 is an approximation of distance versus frequency of the ground at all seasons for night or day. The curve is to all intents and purposes a ground wave suitable for radio compass work. Night effect may be expected. This curve as shown is good for approximate day and night transmission. An increase of power may be required of the lower frequencies for the required distance.

(2) In order to have a well-built structure on which to base allocations, not only for the present, but with a view to taking advantage of future advances in the science of radio, it is desirable that the frequency separation be considered from an international standpoint, and that the same constant of separation, together with the same midpoint for the first channel in each service band, be used by all nations. It would be a simple matter for the United States to allocate the high-frequency channels to

meet national requirements and conditions, but if we must dovetail in with the high-frequency channels of all nations we must use a system common to that of all other nations. What should the separation of the world's high-frequency channels be, and where should we locate the carrier wave for the first channel in each service band?

In most of the governmental services we can today require a 0.1 per cent channel separation for telegraphy, restricting the maximum permissible deviation of frequency from the fundamental to 0.05 per cent of the frequency used, insofar as it applies to apparatus now in use. For a small amount of apparatus under construction we can restrict the deviation to 0.03 per cent, which will permit of a channel separation of 0.06 per cent; and in our Naval Laboratory at Bellevue, D. C., there is a set in operation which has maintained its frequency within 0.01 per cent stability for several months. A few commercial stations in the United States can do as well, if not better.

But it will be several years before even the Navy will be able to equip its many stations with apparatus capable of 0.01 per cent stability, and at least two years before we will be equipped even to use 0.05 per cent separation. A parallel situation exists in the most up-to-date commercial organizations. The average of the less up-to-date commercial stations are not equipped to maintain a frequency stabilization of even 0.1 per cent today, and it will require at least a year before such stations can obtain, install, and learn to operate within an accuracy of 0.05 per cent, or within a 0.1 per cent channel.

The same is true in other countries. All nations, in the International Radio Convention, bound themselves to use modern equipment, but a reasonable time must be allowed to comply with this agreement. Some will be prompt in complying, others will be slow. If one nation uses a 0.2 per cent separation and another 0.5 per cent separation, on adjacent channels, the signals will interfere and business will be interrupted. Consequently, we must build a world-wide allocation structure, and all nations build together.

Exchange of correspondence between various leading technical experts of the world indicates that perhaps 0.2 per cent separation is the best we can hope for for a year or two, and that we may then divide each channel into two, making two channels where one formerly existed; still later on as nations use more accurate control we may again expect that a further halving will become possible. 0.1 per cent separation of channels after a year or two, with allocation of alternate channels for the present, appears a simple and logical solution. For telegraphy it should be practicable to allocate all the 0.1 per cent channels within a year or so. For telephony and facsimile transmission it may not be possible to fill in the alternate channels, at least for some years. Whatever we do, we must play safe, and not build up a structure which will permit us to drift into a situation parallel to the present broadcast situation, where too many stations have been licensed.

(3) The rules for separation and stability of frequency must be capable of enforcement. Otherwise interference will result. Even though stations might be equipped within a few weeks to maintain a frequency stability of 0.05 per cent, what guarantee is there that this will be maintained? The only assurance is through proper inspection by duly constituted authority. The Radio Inspectors do not yet have the necessary modern equipment, nor have they adequate personnel, to cope with such a refined separation, and it will take at least a year before the necessary facilities are available. And this will be equally true in all nations. The traffic police must be procured, must be trained, and must be equipped for the proper performance of their duties.

(4) Assuming that the world is, at the present moment, capable of using channels spaced 0.2 per cent apart without serious interference, an endeavor to bring about an agreement between nations on this point must be sought. This will take time, but it is hoped that this can be done, and that we have sufficient expectation of such agreement that we will be able to license stations without waiting for such agreement. If we take this chance in advance of international agreement, those that are licensed must risk a change in frequency later, if not the loss of such frequency.

(5) It seems desirable that each channel width assigned be the per cent separation multiplied by some mean frequency of a considerable band, rather than the per cent separation multiplied by the exact frequency of the individual channel. The former is simpler, and avoids fractions. I accordingly propose herewith certain rounded values of channel width throughout the high-frequency spectrum, for approximately 0.1 per cent separation. The channel widths for approximately 0.2 per cent separation would be twice the values given.

TABLE II

Frequencies (kc)	Channel width (kc)	Number of Channels	Maximum width (per cent)	Minimum width (per cent)
1500–2198	2	350	0.133	0.091
2200–3313	3	372	0.136	0.091
3316–4400	4	272	0.121	0.091
4405–5490	5	218	0.114	0.091
5495–8202.5	7.5	362	0.136	0.091
8210–10980	10	278	0.122	0.091
10990–16405	15	362	0.136	0.091
16420–21960	20	278	0.122	0.091
21980–23000 (on to 32810)	30	35	0.136	0.130

The frequencies given in Table II are the first and last assignable channel in each of the ranges of frequency. The plan is so arranged that the channel widths progress numerically throughout the entire spectrum without break or skip. With Table II the whole system of assignable frequencies can be built up by anyone, without the necessity for charts or further details.

(6) An analysis of the frequency bands specified in the International Radio Convention is desirable. As Table II shows, there are 2527 channels of the widths given (approximately 0.1 per cent) between 1500 and 23,000 kc. These are divided as follows:

Below 6000 kc,	568	channels in mobile, amateur and broadcast bands
	711	channels in fixed and shared fixed-mobile bands.
Total,	1279	
Above 6000 kc,	368	channels in mobile, amateur and broadcast bands
	880	channels in fixed and shared fixed-mobile bands.
Total,	1248	

Consider the bands assigned to fixed services. We have the following:

1715–2000	Shared with mobile and amateur services. Nothing available for fixed, as amateurs have priority after mobile services.
2000–2250	Shared with mobile services. Of national value for short distance (100 miles) work.
2750–2850	Fixed services.
2850–3500	Shared mobile and fixed services.
3500–4000	Shared mobile, fixed, and amateur. Nothing available to fixed services.
4000–5500	Shared mobile and fixed services.
5700–6000	Fixed services.
6675–7000	Fixed services.
7300–8200	Fixed services.
8550–8900	Shared mobile and fixed services.
8900–9500	Fixed services.
9600–11000	Fixed services.
11400–11700	Fixed services.
11900–12300	Fixed services.
12825–13350	Shared mobile and fixed services.
13350–14000	Fixed services.
14400–15100	Fixed services.
15350–16400	Fixed services.
17100–17750	Shared mobile and fixed services.
17800–21450	Fixed services.
22300–23000	Shared mobile and fixed services.

As shown above, there are a total of 880 channels of the widths given (approximately 0.1 per cent) above 6000 kc in the fixed and shared fixed-mobile bands. Over half of these channels are already in use; between 200 and 250 remain for assignment by all nations. For a basis of approximately 0.2 per cent wide channels, these figures would be cut in two.

The bands assigned to mobile services are the shared bands shown above plus the following:

1500–1715	Mobile	services
2250–2750	"	"
5500–5700	"	"
6150–6675	"	"
8200–8550	"	"
11000–11400	"	"
12300–12825	"	"
16400–17100	"	"
21550–22300	"	"

There is a total of 398 channels available for mobile services at 0.1 per cent separation (alternate channels), of which more than 200 remaining channels are still available in the mobile bands, and only about 30 applications for mobile channels by United States interests, so there will be a little difficulty in finding channels for the applicants. Each mobile station will require at least a day channel and a night channel. From two to ten channels will have to be reserved for each of certain types of mobile services in these bands, for example: short-range portable stations for mining and oil-well investigations, for detecting criminals, for railway services, in the 2250–2750 kc band, and three or four radio-telephone channels for maritime uses with a similar number for use between aircraft and ground stations in each of various bands.

Broadcast Bands. These are as follows:

6000–6150
9500–9600
11700–11900
15100–15350
17750–17800
21450–21550

On the basis of 0.2 per cent separation there are 40 channels, of which 20 channels are available, to be divided between all nations. These bands were allotted by the International Radio Conference primarily for relay telephone broadcasting, or for broadcasting beyond the range of national broadcast networks. It will undoubtedly be more difficult, if at all possible, to increase the number of channels in these bands by lessening the separation in later years, because they will be used mostly for radio telephony.

It may be desirable to adopt a different channelling system for these bands, the 10 kc broadcast channel standard, or a multiple thereof, say 40 kc channels for the present. Owing to the limited number of channels available, these will probably have to be restricted to stations of maximum power, say 5 kw, (or greater), stations which can guarantee reception of high-class programs in foreign countries.

Amateur bands. These are as follows (plus shared bands shown under fixed service list):

7000–7300 kc	
14000–14400	
28000–30000	Amateurs and experimental
56000–60000	Amateurs and experimental

It is expected that the amateur and unreserved bands above 23,000 kilocycles will permit of pioneer experimental work by the amateurs, that they may develop the unknown. Possibly the amateurs will desire to experiment with television and picture transmission in one of their bands. The 1715–2000 band is allocated to mobile, fixed, and amateur service. It is believed desirable to allocate this band to the amateurs.

(7) First attention should be given to the mobile bands, as ships depend on radio and deserve first consideration. Fortunately, priority does not need to be considered within these bands now, as there are plenty of channels. For ships using frequency, one thing must be settled and that is, what should be the common calling frequency? Ships far apart, belonging to the same company, will undoubtedly use their working wave for inter-communication, but those of different companies will often wish to establish schedules, and it must be known in advance what will be the calling channel. There is a mobile band 5500–5700 which is just below the day skip-distance point. This seems a logical band to choose a tentative calling frequency from, and its second or third harmonic can be used when very long distance communication is desired.

High frequency will enable ships of the world to maintain a constant touch with headquarters, at reasonable cost, and this will reduce considerably the overhead expense of shipping.

(8) After the mobile bands, attention should be given to the fixed service bands. The channels, frequency less than 6000 kc, can wait for thorough consideration, as these will not interfere across the seas, and we can easily arrange with our neighboring

North American nations with regard to these channels. Perhaps we can require narrower channels rather earlier for these frequencies.

The channels above 6000 should be assigned without delay, so we can occupy our share of world channels, and have these duly registered in the International Bureau at an early date. The United States should occupy only a certain percentage of the total high-frequency channels, as these channels are for use by all nations. If we appropriated more than our reasonable share it would not make for good feeling between nations.

The channels above 6000 should first be allocated as clear channels, assigning adjacent channels to the same organizations when practicable. By so doing provision is made for later permitting the use of the "band" system, and it may become possible to permit some organizations to fill in the vacant alternate channels, and later again subdivide, sooner than other agencies which are not so progressive. And later, additional stations might be permitted to use some of the same channels on divided time basis and otherwise share frequencies as conditions make this possible. It will be practicable to use some of the high frequencies for day uses not possible at night, and to make special double uses of the so-called summer and winter day frequencies, to considerable advantage.

Also, what about keeping certain bands in harmonic relation? It will be an economic advantage to those who must pay for equipment, if the day and night channels they are to use are in harmonic relation. But this is not always practicable, owing to the fact that the bands are not all in equal harmonic relation. It can be done in most cases.

(9) The question of priority among applicants is most important in the fixed service bands. The following is one suggestion:

Fixed Service Bands.

1. Government—national defense (including coast guard).
2. Government—civil airways.
3. Emergency communication systems, for emergency use only (including provision for necessary testing).
4. Systems for communicating where landwire service impracticable or cannot be secured and maintained without *prohibitive* expense.
5. Trans-oceanic and intercontinental radio service communication systems open to general public service.

6. Certain important experimental and development work.
7. Private trans-oceanic and intercontinental radio communication for very important purposes of national interest, not well served by public service radio companies. (Certain of these may be considered of such importance as to take priority over some under priority No. 5 if no channels remain for allocation to priority No. 5.)
8. National (within the U. S.) radio communication, by radio companies, for public service in competition with wire lines.
9. National (within the U. S.) radio communication by non-radio public service companies, for limited uses by special interests, which could be handled by wire or by public service radio companies.

Another suggestion is to follow the above list, except that Priority No. 8 should be moved up to a place between Priorities No. 5 and No. 6; still another suggestion is to place the press agencies, handling news for all the public, ahead of Priority No. 5. Both suggestions have great merit, and are worthy of serious consideration.

One guiding principle should be that the channels above 6000 kc should be assigned for those who require them to work long distances (international) rather than short distances (national). For example, circuit New York to Japan should be given choice of frequency as compared with the circuit New York to London, whereas the latter should take priority ahead of New York to San Francisco. Considering this rule, together with the fact that there are not enough channels remaining to be allocated, it will be seen that most of those desiring circuits between cities within the United States will have to be satisfied with frequencies below 6000 kc.

It appears likely that there will not be enough channels to accommodate even all of the immediate requirements of the public service radio companies under Priority No. 5; therefore, there will be *considerable opposition* to suggested priority due to the desires of those in Priority No. 7 to obtain channels. The argument then becomes a question of whether the public service common carrier systems should have priority over certain very important private interests not in the public service business. There is much to be said on both sides.

Those who are in the public radio service communication business will say that they have risked their money in pioneering the establishment of radio systems, building up patents through expensive research, training their personnel, arranging their traffic agreements, all for use by the general public, and that if insufficient channels are available all should be assigned to the radio companies as they operate in the service of all the public. Some private interests (such as the press newsgathering organizations) will say that they have not been able to get the radio companies to provide for their needs, and that they will operate in the public interest just as much as the radio companies, if not more so, even though not open to general public service traffic, so they want to maintain their own radio services. The radio companies will reply that they have been so busily engaged in applying their rapid developments that they have not yet had time to satisfy everyone. Personally, I feel sympathetic with all interests, and believe as long as there is going to be competition between at least two great public service communication concerns, the best thing to do (considering that there are far too few channels available) may be to give these companies a large share of the channels in trust, if they demonstrate themselves to be in position to provide adequate service, let competition be their spur, and if the rates are not satisfactory, it may become a matter for regulation by the Interstate Commerce Commission. Or if the service is not adequate and satisfactory, the channels can be taken away from the radio companies and given over to special interests. However, it is a little too early to form a definite conclusion on this subject, and it will be seen that it requires serious consideration. There are, of course, exceptions to all rules, and it may be to the public interest to assign certain of the channels in Priority No. 5 to private interests operating for the benefit of all the public.

(10) Places must be found in the spectrum for other uses of radio. For example, where in the spectrum will television be assigned? For television, channels between 80 and 500 kc width will be required. How soon will television be so reasonably obtainable that the public will demand clear channels for the service? If the time is near at hand, should we not reserve a large space in the spectrum for television development work?

What is to become of the experimenters if all the channels are to be assigned for traffic? Most of the applications from "short-wave" experimenters are for a "broad band," 5000–20000 kc, or something similar. If a station is licensed to use a certain channel it is the duty of the Government to keep others clear of that channel during the term of his license, yet this would be impracticable if experimentation were permitted without restriction. For this reason, experimenters in the future will ordinarily have to be content with a few (not over six) channels, about equal distances apart in the spectrum, and get along as best they can in these bands. Of course, there are exceptions to all cases, and if the experimental work in new fields by suitably equipped experimenters is sufficiently important special experimental channels must be arranged for.

(11) The International Convention requires that modern apparatus be used. "Modern apparatus" must be construed to include limitation of harmonics from all (including broadcast) transmitters, otherwise traffic on high-frequency channels will be interrupted. The rules concerning specifications for apparatus must therefore be decided upon.

In conclusion, the problem of licensing stations to use high-frequency channels is a most difficult one, and the Federal Radio Commission will have much difficulty in building a satisfactory allocation structure to insure the best interest of the public.

Volume 16, Number 9 *September, 1928*

LONG-WAVE RADIO RECEIVING MEASUREMENTS AT THE BUREAU OF STANDARDS IN 1927*

By

L. W. Austin

(Laboratory for Special Radio Transmission Research, Bureau of Standards, Washington, D. C.)

THE monthly averages of the daylight signal intensities of a number of stations and the strength of the atmospheric disturbances, as measured in Washington, are shown in the following tables. Some of the transatlantic stations of the Radio Corporation of America situated at moderate distances from Washington, 250–700 km, have been added to the long-distance stations covered in former reports, since it has been found that signals from stations at such distances are especially useful for comparison with meteorological, solar and magnetic

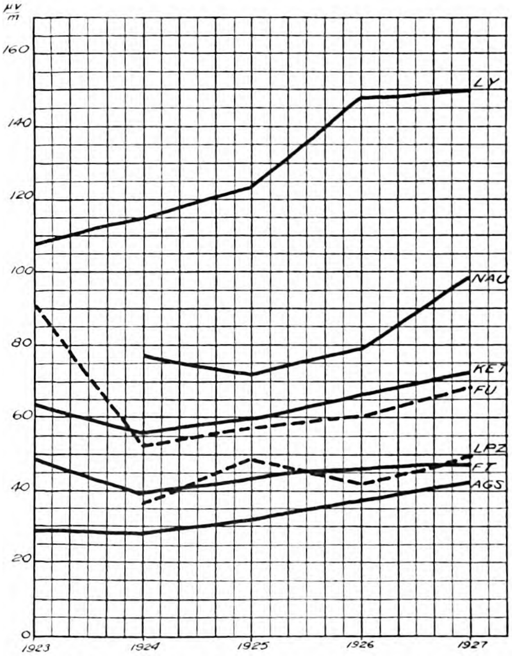

Fig. 1—Annual Average Signal, 10 A.M.

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[*] Original Manuscript Received by the Institute, July 10, 1928. Publication approved by the Director of the Bureau of Standards of the U. S. Department of Commerce.

phenomena. The signals marked A.M. in the tables are received in Washington between 10 and 11 o'clock in the forenoon (E.S.T.) and represent transmission under daylight conditions along the whole wave path except in the case of the two stations, AGS and AGW, near Berlin, where for a period of about six weeks near the end of the year the sun sets slightly before 10 A.M. (E.S.T.). The P.M. signals of the tables, received in Washington between 3:00 and 4:00 o'clock in the afternoon, represent all daylight transmission paths for the southern and western stations measured, and paths partly in darkness and partly in daylight for the European stations except for GBR (Rugby) and GBL (Leafield), which give all daylight transmission during the longest days of summer..

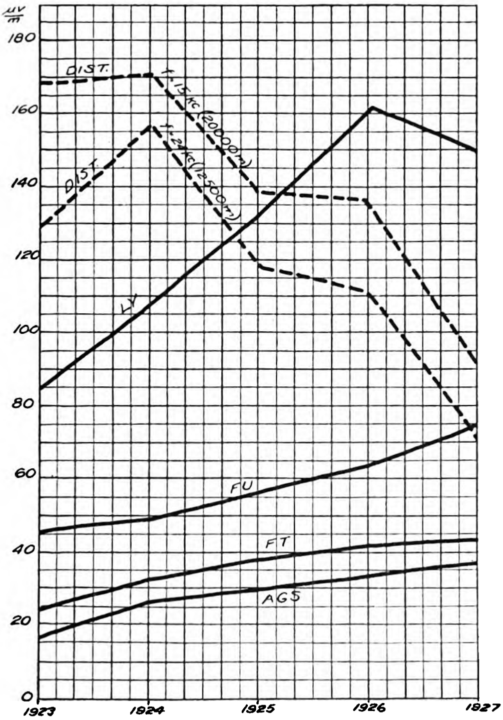

Fig. 2—Annual Average Signal and Atmospheric Disturbances, 3 P.M.

In examining the tables it will be noticed that despite the fact that from January to May 1927 the intensities of the distant stations fell for the most part below the 1926 values, the averages

was less than one half of the 1926 value. The weak disturbances of the summer were undoubtedly connected with the large number of cool days and small number of thunderstorms. There is also a very possible connection between the disturbance weakness and the present high solar activity.

In the figures, the curves of the annual values of signal intensity and atmospherics since 1923 which have been given in the more recent annual reports of the laboratory have been continued to include 1927.

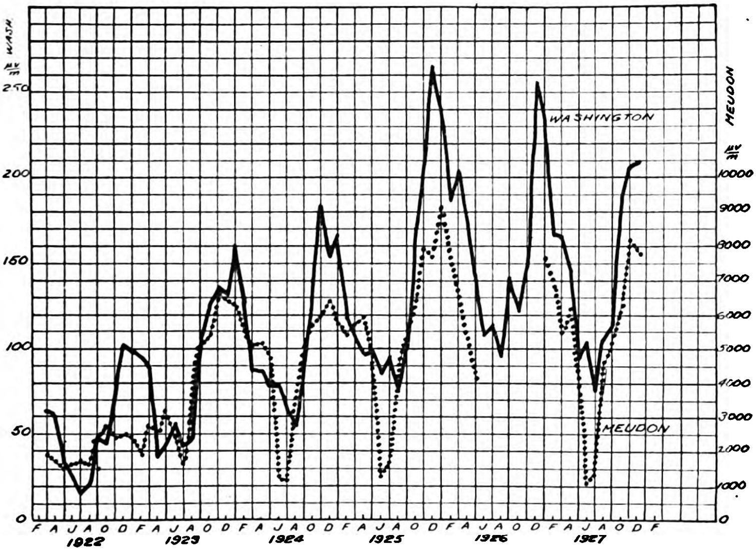

Fig. 4—Lafayette (LY) Average Signal at Washington and Meudon, 3 P.M.

Figs. 1 and 2 show the course of the annual averages of signals from a number of stations as received in Washington in the morning and in the afternoon since 1923. If we except Buenos Aires (LPZ)—and Cayey (NAU), which have been irregular, possibly on account of changes in transmitting current or antenna arrangement, and Bordeaux (LY) in 1927, all signals have increased regularly in intensity from 1924 to 1927 inclusive. The morning signals of Bordeaux have nearly the same observed strength in 1927 as in 1926, while the afternoon signals which have a transmission path partly in darkness have dropped perceptibly in 1927. This behavior of Bordeaux is perhaps due to the fact that the sunspot curve may have now reached its maximum which in the present cycle appears to be less sharp than the maximum in 1917. It now appears from the observations thus far taken in 1928 that the other long-wave stations are following the lead of Bordeaux and are also beginning to fall in strength.

The curves of the atmospheric disturbances in Fig. 2 show, in contrast with the increase in signal, a steady drop until in 1927 the average disturbance strength is only approximately half as great as in 1924.

Fig. 3 gives the seasonal signal variations of various stations for 1927. The autumn maxima usually shown in September by many stations are less definite than in 1926 and in some cases are shifted to October or November.

Fig. 4 continues the comparison of the monthly averages of the U.R.S.I. signals sent out by Bordeaux at 19:55 (7:55 P.M.) G.M.T. as received at Meudon near Paris and at Washington. The figure shows a marked drop in the December signal strength of 1927 from the high values of 1925 and 1926.[1]

The work on the relation of solar activity to radio transmission described in the report for 1926 has been continued and has been printed in another paper.[2]

The signal intensity recorder[3] and the atmospheric direction recorder mentioned in the last report are now in operation and some of the results obtained from them will be published in the near future.

TABLE I
TRANSMISSION DATA, 1927.

	Frequency f	Wavelength λ	Antenna Current I	Effective height h	Distances d km
LY,[4] Bordeaux	15.9	18,900	540	180	6160
FU, Ste. Assise, Paris	15.0	20,000	475	180	6200
FT, Ste. Assise, Paris	20.8	14,400	340	180	6200
AGW,[4] Nauen, Berlin	16.5	18,100	442	170	6650
AGS,[4] Nauen, Berlin	23.4	12,800	389	130	6650
GBR,[4] Rugby	16.1	18,600	600	185	5930
GBL,[4] Leafield	24.4	12,300	195	75	5900
LPZ, Monte Grande, Buenos Aires.	23.6	12,700	600	143	8300
LPV, Monte Grande, Buenos Aires.	17.0	17,600	585	143	8300
SPR, Rio de Janeiro	13.8	22,000	780	150	7800
KET,[4] Bolinas, San Francisco	22.9	13,100	640	51	3920
ICC, Coltano, Pisa	19.9	15,000	380	150	7100
NAU, Cayey, Porto Rico	33.8	8,870	150	120	2490

[4] Daily antenna current reported. Other antenna currents more or less uncertain.

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[1] Author's note: Measurements were discontinued at Meudon during a part of 1926.

[2] PROC. I. R. E., **16**, p. 166; 1928.

[3] PROC. I. R. E., **16**, p. 666; 1928.

TABLE II

Average Signal Intensity and Atmospheric Disturbances for Lafayette (LY), Rugby (GBR), Ste. Assise (FU), Nauen (AGW), Monte Grande (LPV) and Rio de Janeiro (SPR), in Microvolts per Meter.

	A.M.							P.M.					
1927	LY	GBR	FU	AGW	LPV	SPR	Dist.	LY	GBR	FU	AGW	SPR	Dist.
Jan.	147	—	—	60	34	46	26	231	297	97	82	31	37
Feb.	120	110	55	51	45	32	44	163	226	94	63	22	59
Mar.	150	152	66	71	43	34	37	162	182	76	72	27	53
Apr.	171	178	59	73	32	27	47	147	166	72	59	14	67
May	140	154	64	60	26	27	44	95	121	52	42	15	113
June	170	194	89	79	35	37	38	102	117	54	47	19	68
July	160	200	83	78	39	35	50	76	75	46	38	—	310
Aug.	180	198	82	75	34	32	44	105	128	56	51	—	173
Sept.	171	194	76	72	30	30	39	112	138	54	54	—	124
Oct.	174	197	61	72	36	—	31	188	204	89	74	—	44
Nov.	111	128	67	58	—	29	27	207	203	96	81	15	37
Dec.	109	136	60	62	—	32	25	208	230	109	91	—	26
Ave.	150	168	69	68	35	33	38	150	174	75	63	20	93

TABLE III

Average Signal Intensity and Atmospheric Disturbances for Coltano (ICC), Ste. Assise (FT), Bolinas (KET), Nauen (AGS), Monte Grande (LPZ) and Leafield (GBL) in Microvolts per Meter.

	A.M.							P.M.						
1927	ICC	FT	AGS	KET	LPZ	GBL	Dist.	ICC	FT	AGS	KET	LPZ	GBL	Dist.
Jan.	26	34	33	61	38	16	19	40	49	46	66	31	16	28
Feb.	27	27	25	64	39	10	32	35	35	31	54	29	12	44
Mar.	40	40	38	82	51	19	28	38	46	39	69	42	16	43
Apr.	42	44	43	72	41	21	35	24	33	28	50	25	11	47
May	38	43	39	63	34	21	36	20	25	22	47	17	15	90
June	51	54	48	86	50	24	30	23	28	23	54	24	14	56
July	51	55	50	75	46	26	43	—	24	26	43	26	13	230
Aug.	—	54	52	72	51	29	34	—	33	26	48	23	22	130
Sept.	—	62	55	76	50	28	30	—	39	32	54	—	—	93
Oct.	—	65	62	98	56	36	24	—	58	52	76	—	23	41
Nov.	—	42	26	68	66	22	20	—	64	45	75	—	20	30
Dec.	—	44	35	63	57	19	19	—	73	58	85	—	26	20
Ave.		47	42	73	48	23	29		42	36	60	27	17	71

TABLE IV

Average Signal Intensity and Atmospheric Disturbances for Cayey (NAU) in Microvolts per Meter.

	A.M.		P.M.	
1927	NAU	Dist.	NAU	Dist.
Jan.	79	12	63	15
Feb.	79	15	59	20
Mar.	124	15	88	21
Apr.	107	20	73	47
May	82	23	66	45
June	113	18	85	39
July	120	26	—	109
Aug.	94	18	61	85
Sept.	95	22	75	64
Oct.	97	14	97	24
Nov.	111	13	94	19
Dec.	80	13	79	12
Ave.	98	17	76	42

TABLE V

Average Signal Intensity for Rocky Point, L. I. (WSS), Marion, Mass. (WSO), New Brunswick, N. J. (WII) and (WRT), in Millivolts per Meter.

	A.M.				P.M.			
1927	WSS	WSO	WII	WRT	WSS	WSO	WII	WRT
Jan.	4.1	1.0	4.0	—	3.6	0.9	3.4	—
Feb.	3.5	—	3.1	—	3.4	—	3.1	—
Mar.	3.1	1.2	2.8	—	3.2	1.1	3.0	—
Apr.	3.1	1.1	2.3	—	3.1	1.0	2.6	—
May	2.7	0.9	2.2	—	2.8	0.9	2.4	—
June	3.0	1.1	2.5	2.9	3.1	1.0	2.6	2.9
July	2.6	1.1	2.3	2.5	2.5	0.9	2.2	2.4
Aug.	2.7	1.1	2.6	2.6	2.7	0.9	2.5	2.6
Sept.	2.7	1.2	2.4	2.9	2.7	1.1	2.7	3.0
Oct.	3.1	1.3	3.0	3.2	3.0	1.4	3.4	3.4
Nov.	3.9	—	4.2	4.0	3.5	1.4	3.8	3.8
Dec.	2.9	—	3.4	3.3	3.0	1.2	3.3	3.2
Ave.	3.1	1.1	2.9		3.1	1.1	2.9	

Volume 16, Number Number 9 *September, 1928*

MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE*

THIS is a monthly list of references prepared by the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The various articles listed below are not obtainable from the Government. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R000. RADIO COMMUNICATION

R007 Lemoine, S. Allocation of European broadcast wavelengths. *Experimental Wireless* (London), 5, pp. 386–96; July, 1928.

(Discusses basis of reallocation of frequencies to conform with Washington Conference.)

R100. RADIO PRINCIPLES

R113 Barfield, R. H. Woods and wireless. *Nature* (London), 121, p. 908; June 9, 1928.

(Absorption of wireless waves by trees.)

R113 Hankey, H. A. Empire broadcasting—Results of the recent range tests conducted by the Wireless League. *Wireless World and Radio Rev.*, 22, pp. 695–96; June 27, 1928.

(Observations made on 5SW broadcasts during a voyage to the Antipodes via So. Africa. Results showed good reception of this station on 24 meters.)

R113.4 Tuve, M. A. and Dahl, O. A transmitter modulating device for the study of the Kennelly-Heaviside layer by the echo method. PROC. I. R. E., 16, pp. 794–98; June, 1928.

(A method of modulating a transmitting set by sudden pulses of plate current which occur in an unbalanced multivibrator circuit is described.)

R113.5 Pickard, G. W. Some correlations of radio reception with atmospheric temperature and pressure. PROC. I. R. E., 16, pp. 765–72; June, 1928.

(Night reception and temperature at receiving station found to be directly related, maximum reception being associated with maximum temperatures. Temperature effect is local to the receiver. Correlation between night reception and pressure found.)

R113.7 Espenschied, L. Technical considerations involved in the allocation of short waves; frequencies between 1.5 and 30 megacycles. PROC. I. R. E., 16, pp. 773–777; June, 1928.

(Chart given which shows approximate relation of optimum frequency to distance in short-wave radio transmission and the number of available channels based on present practice.)

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* Original Manuscript Received by the Institute, July 14, 1928.

R120 Wilmotte, R. M. The distribution of current in a transmitting antenna. *Jnl.* I. E. E. (London), **66**, pp. 617–627; June, 1928.

(Investigation conducted to ascertain whether assumption usually made regarding distribution of current in antenna is an approximation to conditions obtained in practice. Current distribution of straight vertical antenna found by placing ammeters at various points along antenna. Ammeters placed in inside of cage type of antenna to determine effect of ammeters. Comparison of experimental and theoretical curves showed good agreement over range of 15 to 800 meters.)

R125.6 Chireix, M. Un systeme Francais d'emission a ondes courtes projetees. (A French system of the transmission of projected short waves.) *L'Onde Electrique*, **7**, pp. 169–195; May, 1928.

(Short-wave beam system installed by French government for communication with the colonies is described.)

R125.6 Yagi, H. Beam transmission of ultra short waves. Proc. I. R. E., **16**, pp. 715–41; June, 1928.

(Describes experiments at frequencies above 1500 kc. Curves given which show effect of ground and various types of antennas. Describes the use of the magnetron for the production of 12 cm waves. Circuit arrangements used with these tubes given.)

R127 The equivalent inductance and capacity of an aerial with inserted tuning coil or condenser. *Experimental Wireless* (London), **5**, pp. 357–60; July, 1928.

(Editorial pointing out error in treatment of subject by Palmer in "Wireless Principles and Practice." Mathematical discussion of the conditions.)

R134 Nelson, J. R. Detection with the four-electrode tube. Proc. I. R. E., **16**, pp. 822–39; June, 1928.

(Mathematical analysis of plate rectification and results applied to screen-grid tubes.)

R134.45 David, P. La super-reactions. (On Super-regeneration.) *L'Onde Electrique*, **7**, pp. 217–260; June, 1928.

(Principles of super-regeneration explained. Circuits given. Short bibliography included.)

R141 Bird, L. T. Reactance and admittance curves applied to tuned circuits with and without resistance. *Experimental Wireless* (London), **6**, pp. 371–377; July, 1928.

(Continuation of paper in June issue. Use of vector diagrams in study of circuits.)

R200. Radio Measurements and Standardization

R210 Worrall, R. H. and Owens, R. B. The Navy's primary frequency standard. Proc. I. R. E., **16**, pp. 778–793; June, 1928.

(Describes the method used to determine the fundamental frequency in terms of Naval observatory time to an accuracy of one part in 100,000.)

R214 Van Dyke, K. S. The piezo-electric resonator and its equivalent network. Proc. I. R. E., **16**, pp. 742–64; June, 1928.

(Theory of piezo-electric and mechanical behavior of quartz resonators. Rods which are excited lengthwise through transverse piezo-electric effects and those excited through longitudinal piezo-electric effects are described and equivalent circuits of these are given.)

R261 Turner, H. M. A compensated electron-tube voltmeter. Proc. I. R. E., **16**, pp. 799–801; June, 1928.

(Method described which eliminates source of error in use of electron-tube voltmeter due to changes in filament. The grid bias changes with filament current so the plate current is practically independent of filament current.)

R300. Radio Apparatus and Equipment

R330 Oatley, C. W. The use of a.c. for heating valve filaments. *Experimental Wireless* (London), **5**, pp. 380–384; July, 1928.

(Discusses problem of direct and indirect heating of electron-tube cathode by a.c., concludes that tubes with indirectly heated cathodes have definite place in detector stage but have no obvious advantages in amplifier stages.)

R342 Kirke, H. L. Microphone amplifiers and transformers. *Experimental Wireless* (London), 5, pp. 361–370; July, 1928.

(Design of microphone amplifiers and transformers for use in a broadcast transmitting system.)

R400. RADIO COMMUNICATION SYSTEMS

R431 Tubbs, E. A. System of combatting effects of static. *Experimental Wireless* (London), 5, pp. 378–379; July, 1928.

(Device used by Federal Telegraph Co. of California to increase reliability of service. McCaa system used.)

R500. APPLICATIONS OF RADIO

R522 Kaufman, J. Radio's part in the *Southern Cross* flight. *Radio* (San Francisco), 10, pp. 18–20; July, 1928.

(Description of short-wave transmitting equipment used by the *Southern Cross* on its flight from San Francisco to Australia.)

R526.1 Gunn, R. Aircraft radio and navigation. *Jnl.* Franklin Institute, 205, pp. 849–63; June, 1928.

(Engineering features of aircraft radio and associated difficulties are discussed. Heterodyne beacon system described.)

R526.1 New radio beacon guides planes. *Science and Invention*, 16, p. 338; August, 1928.

(Work of Bureau of Standards on visual type of indicator, and beacon transmitting set at College Park, Md.)

Review of Current Literature*

Prepared by

STUART BALLANTINE

Boonton - - - - - - New Jersey

THE CHIREIX-MESNY DIRECTIVE ANTENNA FOR SHORT WAVES

UNDER this caption may be reviewed some interesting and novel developments which have lately been carried out in France in the field of directive radiation and reception at short wavelengths. Among the contributors to these developments the names of H. Chireix, Chief Engineer of the Societé Francaise Radioelectrique, and René Mesny, Professor of Hydrography in the French Navy, are conspicuous. I am indebted to conversations with Prof. Mesny for an account of this work and to the publications of Mesny and Chireix cited below.[1]

1. Fundamental Types of Directive Antenna Arrays. Fig. 1 illustrates an array which will be designated as of Type 1. Along the x axis is arranged a uniform series of vertical wires, regularly

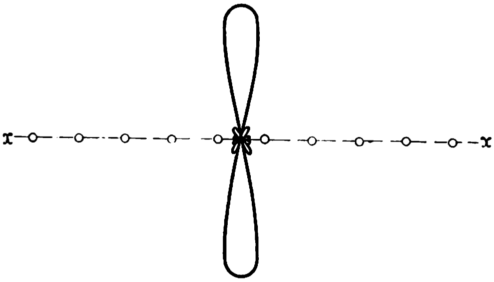

Fig. 1—Directivity Characteristic of Alignment of Type 1; Currents in Phase.

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* Original Manuscript Received by the Institute, July 31, 1928.

[1] René Mesny: "Les Ondes Electriques Courtes," p. 76, (Paris, 1927); *L'Onde Electrique*, 6, 181, 1927; *Compte rendu Académie des Sciences*, Report of Meeting of May 2, 1927, 1, 1047; *Tejd. van Het. Neder Radiogenootschap*, 3, Feb. 1927.

H. Chireix: *Radio-Electricité* (Sup. Tech.), 5, 65, 1924; *L'Onde Electrique*, 5, 237, 1926; 7, 169, 1928; French Patent No. 216,757, filed Mar. 10, 1926.

spaced a distance of less than one wavelength, and carrying currents in the same phase. The radiation from this array is principally concentrated in two symmetrical beams along the y axis; and the acuity of these beams increases with the horizontal length of the array. The whole arrangement is equivalent to a hyperbolic reflector of infinite focus. If n represents the number of wires, and d their separation, the electric field intensity at a distant point may be expressed as

$$E = \text{Const.} \frac{\sin \dfrac{n\pi d}{\lambda} \sin \varphi}{\dfrac{n\pi d}{\lambda} \sin \varphi} \tag{1}$$

This has a grand maximum when $\varphi = \pi/2$ and other maxima of negligible importance. As to these small maxima it is perhaps of academic interest to note that Bellini has shown that they may be completely eliminated by making the currents in the separate wires proportional to the Binomial Coefficients; Mesny has also suggested that the most important of them may be reduced from 1/22 to 1/225 of the intensity of the principal maximum by graduating the currents according to the law: $\cos \pi x/2$.

As a second fundamental type of assembly consider Fig. 2. Here the antennas are situated along the y axis and their several currents are dephased by a constant angle corresponding to their angular separation d; viz. $\phi = 2\pi d/\lambda$. Thus if the separation is a quarter-wavelength the currents in adjacent wires will be in quadrature ($\phi = \pi/2$). When the separation is $\lambda/4$ the beam is directed along the y axis and has the remarkable property of being asymmetrical. For a given number of wires and total horizontal length of the alignment the Type 2 beam is broader than Type 1. One recognizes here immediately the close identity in the operation of this system and the "wave directors" of Uda and Yagi.[2]

Types 1 and 2 may be combined in a general two-dimensional array and it can be shown mathematically that the resulting directional characteristic is the product of the characteristics of the separate ranks and columns regarded as Type 1 and Type 2 systems. The outstanding property of curtains of Type 1 is the acuity of the beam; the outstanding property of Type 2 is the possibility of undirectional, or at least asymmetrical, radiation. In practice these types are combined to secure both results. A sufficient asymmetry of the beam is obtained by employing but two of the Type 2 elements, so that the antenna takes the form of two curtains of Type 1 separated by a fraction of a wavelength. The separation which secures the greatest asymmetry is $\lambda/4$, the phase difference between the curtains being then adjusted to $\pi/2$. To secure an acute beam the horizontal dimensions of the curtains are made as large as possible, since the acuity has been shown by Mesny to be proportional to the ratio of this dimension to the wavelength. This applies to the normal or principal beam.

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[2] H. Yagi and S. Uda: *Proc.* Imp. Acad. **2,** No. 2, 1926. PROC. I. R. E., **16,** 715; June, 1928.

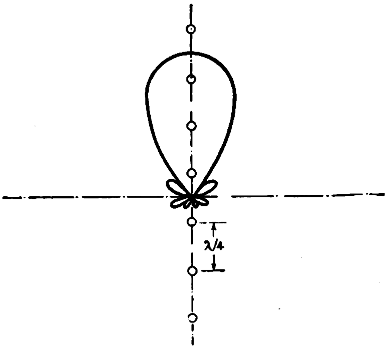

Fig. 2—Directivity of Alignment of Type 2; Currents in Adjacent Wires in Quadrature; Separation of Wires $= \lambda/4$.

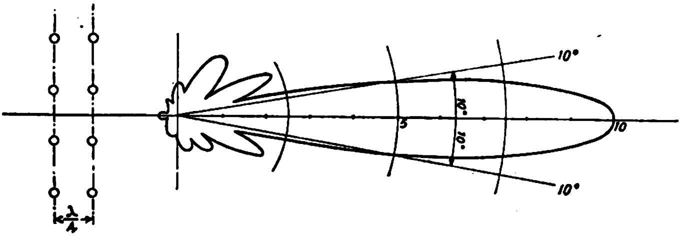

Fig. 3—Directivity of Franklin Projector of Eight Elements As Determined Experimentally at Short Distances (after Mesny).

The amplitude of the other maxima, which may be properly called "diffraction fringes," decreases as the number of vertical wires increases for a given total horizontal length.

2. **Franklin System.** Such a double curtain system as devised by C. S. Franklin of the English Marconi Company is shown in Fig. 3 together with its directivity characteristic.

3. Mesny's Arrangement in the Grecian Key-Pattern. To avoid some of the difficulties of the rather delicate adjustment of current amplitudes and phases in the wires of the preceding antennas and to simplify the problem of supplying power to them, Mesny has proposed a rather ingenious antenna arrangement comprising one wire bent into the form of the Grecian key-pattern. (Fig. 4.) The current distribution is indicated by the dotted lines and the directions of the currents by the arrows. It will be noted that in the vertical sections the currents are all in the same direction. An important advantage which Mesny claims for this system is the uniformity of the flow of power, there being no significant reflection points throughout the length of the wire. The radiation in a vertical direction is negligible at large distances because the currents in each of the horizontal elements are in phase opposition and their effects cancel. The Grecian curtain may be excited in the middle or at the ends by means of separate generators, the currents of which are automatically synchronized by taking advantage of the interlocking action of two oscillators. To secure an undirectional beam two Greek curtains may be mounted with a separation of λ/4 as in the Franklin system; one of these then acts as a "reflector."[3] The distribution of energy in a horizontal plane about such a Grecian antenna has been investigated experimentally by Mesny and is shown in Fig. 5.

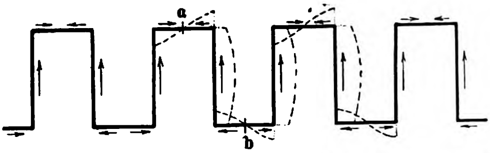

Fig. 4—Directive Antenna Arrangement of Mesny in the Grecian Key-pattern; Current Distribution Shown by Dotted Lines, Directions by Arrows. Distance $ab = \lambda/2$.

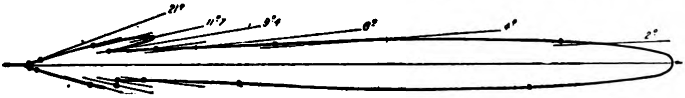

Fig. 5—Directivity of Grecian Curtain (Fig. 4) $\lambda = 5.87$ m. Symmetrical Left-hand Lobe of the Curve Has Been Omitted.

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[3] O. B. Blackwell (*Bell Tech. Journal*, April 1928, p. 179) has recently described a short-wave antenna used at Cliffwood, N. J. for trans-Atlantic telephony which appears to me to be precisely of this Grecian pattern type of Mesny, although it is not specifically ascribed to him.

4. Arrangement of Chireix. Another arrangement devised by H. Chireix and based upon the preceding is shown in Fig. 6. The wire is here arranged *en zig-zag* and the distribution of current is shown by the dotted lines and the directions are indicated by the arrows. Each branch will be seen to comprise a half-wave element. The wire offers no change in characteristic impedance and the energy is readily propagated from one end to the other. The distribution diagram resembles in a general way that obtained with the Grecian curtain, but the amplitudes of the higher order maxima are advantageously smaller. The polarization of the electric field on the horizon is vertical.

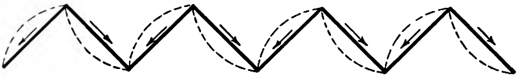

Fig. 6—Zig-zag Directive Arrangement of Chireix, a Modification of Fig. 4.

A practical form of this antenna is shown in Fig. 7. This is obtained by a double reproduction of Fig. 6. Power is supplied by means of the wires *n m*. It will be observed that the wires *aa′ bb′ constitute effectively half-wave elements of the same sign in series,* thus increasing the radiated energy in the zenithal plane. The curtain may be extended vertically by repetitions of this structure, yielding the Chireix-Mesny (C. M.) antenna system illustrated in Fig. 8, which comprises a total of 32 half-wave elements. In the complete antenna a similar curtain is usually employed as a reflector to render the beam asymmetrical. The total length of the curtain is 5 or 6 wavelengths and the beam obtained is less than 20 deg. in breadth. Power is supplied to the middle of each bay by the lines *nn* and *mm*, which are terminated in coupling impedances designed to minimize reflection.

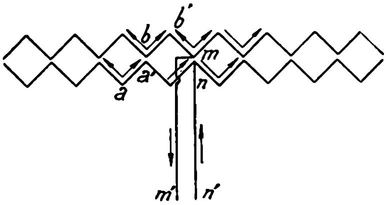

Fig. 7—Practical Form of the Chireix Antenna.

It may be added that, according to Chireix, the adjustment of the system is not critical, but a variation of 200 kc, plus or minus, is tolerable at a wavelength of the order of 25 meters. This is more than sufficient to permit the simultaneous use of several frequencies with the same antenna.

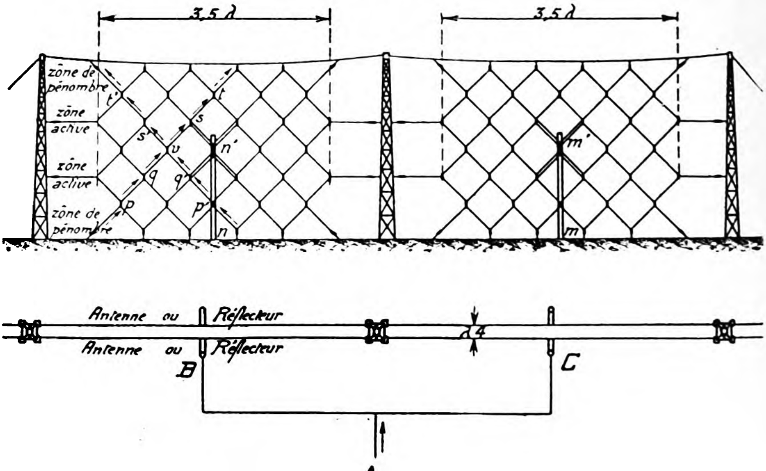

Fig. 8—Design for Complete Chireix-Mesny (C. M.) Projector of Two Bays Comprising 32 Half-wave Elements.

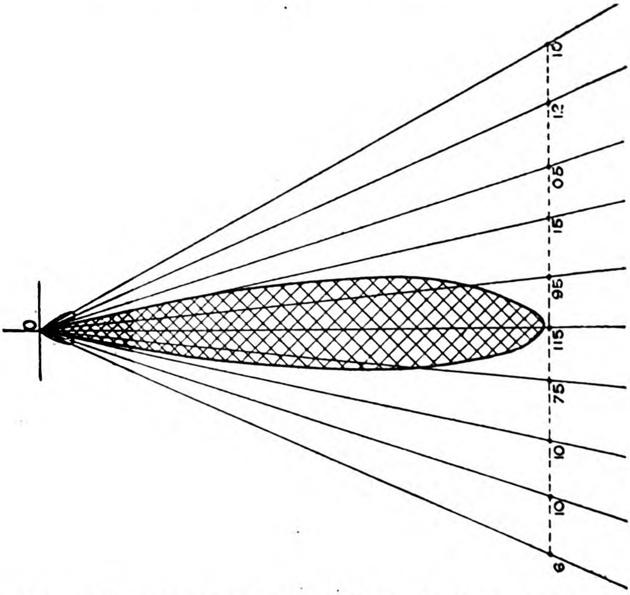

Fig. 9—Directivity of A Single Bay of the C. M. Projector Shown in Fig. 8, As Determined Experimentally at Short Distances (Chireix).

The directivity of a single bay of the C. M. antenna is shown in Fig. 9, which is an experimental curve obtained at *short* distances. The angular width of the beam is about 20 deg. With a reflector the two beams are in the ratio of 1.8 (at short distances). The direction of the beam may be reversed by exciting one or the other of the two antennas. This is an enormous advantage in practice in avoiding multiple ghost-signals due to complete transmission around the earth.

The C. M. antenna is also used for reception and is said to augment the power of the received signal 4 or 5 times over the reception with an ordinary antenna.

5. Installation of the C. M. Projector at Sainte-Assise, France. An installation of the C. M. antenna has been made by the Societé Francaise Radioelectrique in the short-wave station recently constructed by the Companie Radio-France for the French government. This apparently represents the first full-scale test of this system. A single bay of the antenna, as installed at Sainte-Assise, is shown in Fig. 10. This station (FW3) is in regular communication with the Argentine and the Marconi beam station in Brazil. At the wavelength employed (15.5 m), operation with Buenos Aires is possible from 7 A.M. to 10 P.M. The results are considerably better than those obtained with ordinary non-directional systems and compare favorably with those of other beam systems.

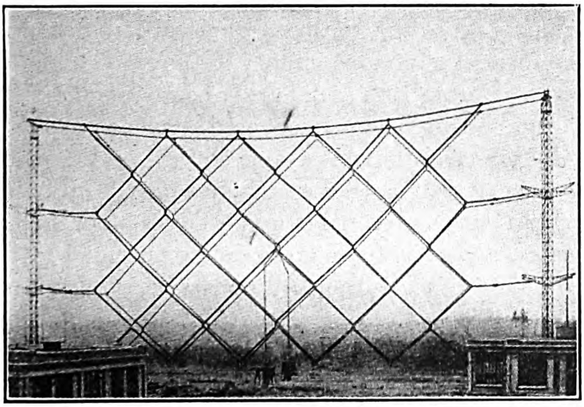

Fig. 10—View of Installation of a Single Bay C. M. Projector at Sainte-Assise, France.

BOOK REVIEWS

Wireless Principles and Practice, By L. S. Palmer. Published by Longmans, Green & Co., Ltd., London, 1928. 492 pages, 307 illustrations. Price $7.00.

This book in a very thorough manner deals with the apparatus and theory of present day radio. After a consideration of radio circuits and their calculation, the theory of the vacuum tube and its many applications are discussed. Chapters on Wireless Telephony and Directional Wireless close the book.

The completeness of the bibliographies at the end of each chapter make it highly desirable as a reference book to the technical man.

C. R. Hanna†

National Physical Laboratory Report for 1927. Issued by H. M. Stationery Office, Astral House, London W. C. 2. 264 pages paper bound. 7s. 6d.

A report of this nature, when prepared by so extensive a research organization as the National Physical Laboratory of England, in the form of a broad summary of the work undertaken by the Laboratory during 1927, can hardly be reviewed satisfactorily in a brief form. Many fields are covered, such as physics, electricity, metallurgy, aerodynamics, and several of the engineering fields; but in nearly all cases the technical details and other data concerning the various problems under advisement are omitted. In many cases, however, these facts can be obtained by referring to the official and other authorized papers which are listed, that have appeared during the year prepared by various members of the Laboratory staff.

A few sections of interest to radio engineers will be referred to; such as Sound, in which research on the acoustics of buildings, effect of partitions, the measurement of the absorbing power of various materials, the absolute measurement of sound intensity, and other acoustical determinations are outlined.

The international comparison of Radio Frequency Standards during the year by means of piezo oscillators (transported to England by the Bureau of Standards) is mentioned. An interesting method is outlined which was developed for the investigation of modes of vibration of flat quartz crystals, utilizing the principle of optical interference. The upper electrode is made somewhat transparent by utilizing a fine copper gauze which has been flattened and partially dissolved in acid. Above this a plate of glass (optically plane) is supported, the spacing being carefully regulated to produce interference bands when viewed with the proper light, when the crystal is not excited. When oscillating the spacing of course varies and the bands become blurred or disappear.

---

† Research Laboratory, Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pa.

The radio section outlines the research being carried out on propagation of waves, directional wireless, interference from transmitting stations, solar eclipse of June 29, 1927, and rotating beacons. For the results of the very extensive tests on errors in direction finding the reader is referred to special report No. 5–1927, by R. L. Smith-Rose.

A cathode-ray oscillograph circuit for measuring wave form at radio frequencies and for other routine tests has been developed, details of which are included in the report. Measurements on a double detection receiver, the circuit for a long-wave time receiver, and the study of current distribution in antenna systems, and details of a short-wave transmitter complete this section. The latter item has a power of 200 watts and covers a range of from 5 to 30 meters. The photograph and circuit details indicate that the push-pull type of circuit, familiar in longer wave oscillators, can be adapted to the 5-meter band. A novel method of modulating this system is shown, in which the plate circuit of a modulator tube is coupled inductively to a coil in the high-voltage supply lead of the oscillator tubes. The modulated note is derived by audio-frequency tuning in the grid circuit, with keying accomplished by shorting out a small inductance coupled to the grid coil (presumably of the modulator circuit).

This review necessarily must omit many items referred to, many of which often consist of but a few sentences.

R. R. Batcher†

---

† Decatur Manufacturing Co., Inc., Brooklyn, New York.

APPLICATIONS FOR MEMBERSHIP

Applications for transfer or election to the various grades of membership have been **received from the persons listed below, and** have been acted upon by the Committee on Admissions. **Members objecting to transfer or** election of any of these applicants should communicate **with the Secretary on or before Septem**ber 31, 1928. These applications will be considered **by the Board of Direction of the Institute at** its October 3rd meeting.

For Transfer to the Fellow grade

Dist. of Columbia	Washington, Navy Department	Hooper, S. C.
Massachusetts	Cambridge, Harvard University	Kennelly, Arthur E.
New York	Schenectady	Rice, Chester W.

For Election to the Fellow grade

Dist. of Columbia	Anacostia, Naval Research Labs	Wheeler, Lynde P.

For Transfer to the Member grade

New Jersey	New Brunswick, P. O. Box 392	Cumming, Kenneth N.
New York	Hollis, 100–23 198 St	Hotopp, Alfred H., Jr.

For Election to the Member grade

Illinois	Chicago, 623 S. Wabash Ave	Smith, K. R.
Maryland	Chevy Chase, 6619 Summit Ave	Walls, H. J.

For Election to the Associate grade

California	Fullerton, 113 E. Commonwealth	Finseth, E. B.
	Los Angeles, 6019 S. Broadway	Newman, William H.
	Monterey Park, 1012 E. Columbia St	Sanmann, Edgar J.
	San Diego, 4741 32nd St	Wright, Robert E.
	San Francisco, 511 Matson Bldg	Attmore, William B.
	San Francisco, 245 Market St	Kellogg, Richard B.
	West Hollywood, 871 Hilldale St	Threlkeld, Howard M.
Colorado	Hereford	Nichols, Fred A.
	Pueblo, 2409 Pine St	Glasscock, Glen R.
Connecticut	Wilson, 8 Garden Street	Appleby, Bertie
Delaware	Wilmington, 2303 Franklin St	Boylan, Brandt
Dist. of Columbia	Washington, Federal Radio Commission	Blackwell, G. C.
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	Chicago, 3843 N. Ridgeway Ave	Sorensen, Carl P.
	Downers Grove, 720 Maple Ave	Johnson, Arthur R.
Massachusetts	Boston, 332 A Street	Baird, Hollis S.
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	East Springfield, 26 Prentice St	Hurff, Jos. L.
	Lowell, 31 Princeton St	Morton, Clarence F.
	Springfield, 491 Wilbraham Rd	Der Bedrossyan, Mark
Michigan	Lapeer, Drawer A	Margraf, Frank J.
New Jersey	Camden, Victor Talking Machine Co	Cooley, Elihu H.
	Hoboken, Cooper Hewitt Elec. Co	Dana, David W.
	Hoboken, 624 Washington St	Izzo, Anthony
	Netcong, P. O. Box 393	Mack, Paul
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	Brooklyn, Y. M. C. A. Hanson Place	Donner, Louis F.
	Brooklyn, 654 East 23rd St	Stobbe, John A.
	Buffalo, 225 Massachusetts Ave	Pfleegor, Carroll M.
	Buffalo, 572 Elmwood Ave	Vanacore, Thomas
	New York City, 66 Broad St	Beltz, Willis H.
	New York City, 237 W. 100th St	Jones, Dramin D.
	New York City, 463 West Street	Koerner, Allan M.
	New York City, 1054 Grant Av. Bronx	Siegal, Jos A.
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	Charlotte, c/o Radio Sta. WBT	Reid, Ralph J.
	Gastonia, c/o A. Kirby & Co	Jenkins, Russell A.
	Gastonia, 413 Broad St	Morris, Robert S.
Ohio	Cleveland, 15305 Huntmere Ave	Melrose, James W.
Oregon	Portland, Reed College, Physics Dept	O'Day, Marcus

Pennsylvania	Philadelphia, 1603 Ellsworth St	Caporate, Peter
	Pittsburgh, 2012 Noble St	Bricker, James O.
	Pittsburgh, 328 Stratford Ave	Wallace, H. E.
	Wilkinsburg, 442 Franklin Ave	Haller, G. L.
Tennessee	Athens, 12 Euclid Ave	Knight, Gaylord A.
Chile	Valparaiso, 98th Tubildad St	Basaure, Auro
China	Tientsin, 10 Recreation Rd	Chen, Ying-Chien
England	Dorset, Long Crickel, Wienborne	Harris, R.
	Essex, 11 Argyll Rd. Westcliff-on-Sea	White, Thomas G.
	Gloucester, 29 Lannett Rd	Myers, Geo. T.
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	Tokyo, Electro Technical Lab	Matsumura, Sadao
	Kumamotoshi, c/o Shimizu Hosojo	Shimayama, Tsurno
Scotland	Glasgow, 105 Douglas St	Carnie, Ben

For Election to the Junior grade

Massachusetts	Marion, c/o R. C. A	Carson, Jos.
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	Minneapolis, 2515 Irving Av. So	Gould, Payson R.
New York	Buffalo, 39 Charleston Av	Patterson, Curtis B.
Pennsylvania	Philadelphia, 1533 Pine St	Gilson, Walter E.
Texas	Amarillo, Route 3, Box 99–B	Reville, T. A., Jr.

PROCEEDINGS OF
The Institute of Radio Engineers

Volume 16 October, 1928 Number 10

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GENERAL INFORMATION

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Payment of the annual dues by a member entitles him to one copy of each number of the Proceedings issued during the period of his membership.

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It is understood that the statements and opinions given in the Proceedings are the views of the individual members to whom they are credited, and are not binding on the membership of the Institute as a whole.

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Dinner for Professor Zenneck

On the evening of September 5, 1928, preceding the regular meeting of the Institute in New York, a dinner was held in honor of Professor Jonathan Zenneck, recipient of the Institute's Medal of Honor for 1928, in the McAlpin Hotel.

Ninety-six members of the Institute attended the dinner, a photograph of which appears as our frontispiece for this month. Dr. Alfred N. Goldsmith made a short speech of welcome to which Professor Zenneck replied.

Seated at the speakers' table, in the right-hand background, are R. H. Marriott, Past-President; A. H. Grebe, J. V. L. Hogan, Past-President; Melville Eastham, Treasurer; L. E. Whittemore, Vice-President; Professor Zenneck, Alfred N. Goldsmith, President; Ralph Bown, Past-President; Donald McNicol, Past-President; R. A. Heising, J. H. Dellinger, Past-President; and John M. Clayton, Secretary.

CONTRIBUTORS TO THIS ISSUE

Andrew, Victor J.: Born August 31, 1902, in Medina County, Ohio. Amateur experimenter and operator since 1919. Received the B.S. degree from College of Wooster, 1926, and the M.S. degree from the University of Chicago, 1928; research section, Signal Corps Radio Laboratories, Ft. Monmouth, N. J., 1926–27; Electric and Manufacturing Company, 1928. Associate member of the Institute, 1925.

Crossley, Alfred: (See PROCEEDINGS for April, 1928).

Graham, Virgil M.: Born January 22, 1902, at Rochester, N. Y. Attended University of Rochester, 1920–23; radio engineering work with Stromberg-Carlson Telephone Manufacturing Company, 1923 to date; for the past three and one-half years in charge of radio engineering laboratory; Chairman of Rochester Section of Institute, 1927; Secretary of Rochester Engineering Society; Associate member, A.I.E.E.; Associate member, Institute of Radio Engineers, 1924; Member, 1927.

Hoag, J. Barton: Born September 22, 1898, in Colorado Springs, Colo. Amateur radio experimenter since 1916. Instructor Army Radio School at Colorado College, 1918, and received A.B. degree, 1920; instructor, including course in radio, at Colorado College, 1921. Assistant in Physics at University of Chicago, 1923; Ph.D. in physics with minor in mathematics at University of Chicago, 1927; 1928 to date instructor, including radio courses. Member, American Physical Society.

Jakosky, J. J.: Born January 20, 1895, at Vinita, Oklahoma. B.S. degree in mechanical engineering, University of Kansas, 1920; B.S. degree in electrical engineering, University of Pittsburgh, 1924; M.E. degree, University of Kansas, 1926. Liaison, radio, and signal officer, U. S. Army, 1917–18; research engineer, U. S. Bureau of Mines, 1920–25; in charge of underground communication investigations and laboratory, U. S. Bureau of Mines, 1924–25; consulting research and development engineer, 1925 to date, including development of interference eliminator for Research Corporation and the Western Precipitation Corporation for elimination of interference caused by Cottrell high-voltage dust collection equipment; consulting engineer in research and development for Radiore Corporation and Southwestern Engineering Corporation. Member of the Institute, 1924.

Jansky, C. M., Jr.: Born at Delton, Michigan. Graduated University of Wisconsin, master's degree in physics, 1919. Since January 1, 1920, in charge of instruction in radio engineering at the University of Minnesota and of operation of that university's broadcasting station. Member of all four National Radiotelegraph Conferences. Associate professor in radio engineering at University of Wisconsin and consulting radio engineer. 1928, member, Committee on Standardization, I.R.E.; Associate member, I.R.E., 1918; Member, 1925; Fellow, 1928.

Page, Robert M.: Born at St. Paul, Minnesota, June 2, 1903. B.S. degree in physics, Hamline University, St. Paul, Minn., 1927. Engaged in radio research and precision measurements, U. S. Naval Research Laboratory, 1927 to date. Associate member, I.R.E., 1927.

Terman, Frederick Emmons: (See PROCEEDINGS for April, 1928).

Wheeler, Harold A.: (See PROCEEDINGS for January, 1928).

INSTITUTE NOTES AND RELATED ACTIVITIES

SEPTEMBER MEETING OF THE BOARD OF DIRECTION

At the meeting of the Board of Direction of the Institute, held at 2:00 P.M. on September 5, 1928, in the offices of the Institute, the following were present: Alfred N. Goldsmith, President; L. E. Whittemore, Vice-President; Melville Eastham, Treasurer; Ralph Bown, Junior Past-President; J. H. Dellinger, R. A. Heising, J. V. L. Hogan, R. H. Marriott, and John M. Clayton, Secretary.

The following were transferred or elected to the higher grades of membership in the Institute. Transferred to the Fellow grade: W. R. G. Baker, Harold Beverage, O. B. Blackwell, Carl R. Englund, R. V. L. Hartley, Fred H. Kroger, Richard H. Ranger, and Owen D. Young. Transferred to the Member grade: William J. Blenheim, James K. Clapp, Francis H. Engel, David Grimes, Harold T. Melhuish, Douglas Rigney, Maurice C. Rypinski, George E. Sterling, Lee Sutherlin, George S. Turner, Paul T. Weeks, and W. A. Winterbottom. Elected to the Member grade: Milton F. Bickford, William M. Browner, W. E. Downey, A. T. Haugh, W. A. Hillebrand, G. Madgwick, Francis E. Pierce, and F. V. Sloan.

One hundred and eleven Associate members and sixteen Junior members were elected.

NOMINATIONS FOR OFFICERS AND MANAGERS FOR 1929

The Board of Direction, at its September 5th meeting, made the following nominations for 1929 officers and members of the Board of Direction:

For President: A. Hoyt Taylor, Laurens E. Whittemore.

For Vice-President: Charles P. Edwards, Alexander Meissner.

For Managers (Three-Year Term): Lawrence C. F. Horle, C. M. Mansky, Jr., William Wilson.

For President

A. HOYT TAYLOR

A. Hoyt Taylor was born on January 1, 1879, in Chicago, Illinois. He was graduated with the B.S. degree from Northwestern University and became an instructor in physics and electrical engineering at Michigan State College in 1900. In 1903 he was appointed an instructor in electrical engineering at the University of Wisconsin. In 1908 he left the University to study at the University of Gottingen, Germany, from which he received the Ph. D. degree. Upon returning to the United States he was elected professor of physics and head of that department at the State University of North Dakota, where he remained until 1917. Dr. Taylor entered the Naval Reserve as a Lieutenant in March, 1917, as district communication officer at Great Lakes Naval Training Station. He was transferred to Belmar, N. J. as transatlantic communication officer in the fall of 1917 and in 1918 was promoted to Lieutenant Commander with headquarters at Hampton Roads. Early in 1919 he was ordered to Washington, D. C. and placed in charge of the aircraft radio laboratory. During this period he was promoted to Commander, remaining in active duty until July of 1922. When the Naval Research Laboratory at Anacostia, D. C. was organized, Dr. Taylor was made Superintendent of its Radio Division, in which capacity he has served to date. He was elected to Associate membership in the Institute in 1916, and was transferred to the Fellow grade in 1920. In 1927 he was awarded the Morris Liebmann Memorial Prize by the Institute.

LAURENS E. WHITTEMORE

Laurens E. Whittemore was born in Topeka, Kansas, August 20, 1892. He was educated at Washburn College, Topeka, Kansas (A.B. degree in 1914) and the University of Kansas from which he received the M.A. degree in 1915. Mr. Whittemore was an instructor in the Department of Physics at the University of Kansas from 1915 to 1917. He left the University to join the Radio Laboratory Staff of the U. S. Bureau of Standards where he remained until 1923. From 1923 to 1924 he was Secretary of the Interdepartment Radio Advisory Committee of the U. S. Government. He served as a member or as secretary of the four National Radio Conferences held by the U. S. Department of Commerce in 1921–1924, and was secretary of the American Delegation to the International Radio Conference held in Washington in 1927. Since 1925 he has been an engineer in the Department of Development and Research of the American Telephone and Telegraph Company in New York City. Mr. Whittemore became an Associate member of the Institute in 1916. He was appointed to the Board of Direction of the Institute in 1926, and served as chairman of the Committee on Standardization of the Institute from 1926 to 1928, inclusive. He is now a Fellow in the Institute and has been Vice-President of the Institute during the year 1928. He is the author of several Bureau of Standards publications and was a joint author of the "Lefax Radio Handbook."

For Vice-President

CHARLES P. EDWARDS

Charles P. Edwards was born on December 11, 1887. He attended the Arnold Technical School at Chester, England, and the Marconi Engineers' Training School at Chelmsford, England. In 1903 he was appointed technical assistant to the English Marconi Company in which capacity he served until 1906, when he became radio engineer for the Canadian Marconi Company. In 1909 he was appointed general superintendent of radio-telegraphs for the Marine Department of the Canadian Government. In 1911 that branch of the Canadian Government was transferred to the newly created Department of Naval Service. He was appointed officer of the "Order of the British Empire" for war services in 1918 and holds the rank of Lieutenant Commander, R.N.C.V.R. Commander Edwards was a Canadian delegate to the International Radiotelegraph Convention in London in 1912, and a delegate for Great Britain to the International Communication Conference in Washington in 1920. From 1921 to 1922 he was Chairman of the Ottawa Branch of the Engineering Institute of Canada of which he is an Associate member. He was elected to the Member grade in the Institute in 1913 and was transferred to the Fellow grade in 1915.

ALEXANDER MEISSNER

Alexander Meissner was born in Vienna, Austria, on September 14, 1883. He was educated at the Vienna Technical School from which he received the degree of Doctor of Technical Science in 1909. In 1922 he received the degree of Honorary Doctor of Engineering from the Technical High School of Munich. In 1925 Dr. Meissner was awarded the Heinrich Hertz gold medal for his inventions in radio and its allied arts. Since 1907 he has been with the Telefunken Gesellschaft, and is at the present date head of the research laboratory of the Telefunken organization. He is an honorary professor at the Berlin Technical High School. Dr. Meissner has patented a great number of radio circuits and appliances and is credited with the invention of the Telefunken compass, musical quenched spark systems, impulse excitation spark gaps, three-electrode tube oscillators, etc. Dr. Meissner has contributed a number of papers to the PROCEEDINGS of the Institute of Radio Engineers as well as other technical journals. He was appointed to membership on the Institute's Committee on Standardization in 1928. In 1914 he was elected to Associate membership in the Institute and in 1915 he was transferred to the Fellow grade.

For Managers Three-Year Term

ARTHUR BATCHELLER

Arthur Batcheller was born at Wellesley, Mass. in 1888. He studied electrical engineering for three years at the Y. M. C. A. Polytechnic School at Boston, Mass., and was instructor for four years in radio and electricity at the evening session of the Boston School of Telegraphy. Mr. Batcheller was assistant to the electrical engineer in charge of installation and maintenance of railroad electric signal system for the Boston, Revere Beach and Lynn Railroad for four years. From 1915 to 1916 he was employed as chief electrician for the Eastern Steamship Corporation, Boston, Mass., and from 1917 to 1919 U. S. Radio Inspector in charge of New England District. In 1919 he was associate partner and founder of the Massachusetts Radio Telegraph School, engaged in training students in electricity and radio. In 1920, Mr. Batcheller accepted an appointment as radio inspector in charge of the 2nd Radio District, with headquarters in New York, and has served in that capacity to date. In 1925, he received the appointment of Lieutenant Commander, CV (S), U. S. N. R. Mr. Batcheller is a member of the Committee on Conference to study safety of life at sea, to be held in London in 1929. He was elected to Associate membership in the Institute in 1914, and transferred to Member grade in 1920.

LAWRENCE C. F. HORLE

Lawrence C. F. Horle was born in Newark, N. J. on May 27, 1892. He was educated in the primary, grammar, and high schools of Newark, N. J. and at Stevens Institute of Technology of Hoboken, N. J., from which he received the M.E. degree in 1914. From 1914 to 1916 he was a member of the faculty of the Department of Physics at Stevens Institute, during which period a large part of his time was spent in investigations of vacuum-tube characteristics and development of vacuum-tube circuits. From 1916 to 1917 Mr. Horle was designing engineer of the Public Service Corporation of Newark. During the war (from 1917 to 1920) he was Expert Radio Aid, Navy Department, in charge of the radio development laboratory at the Washington Navy Yard. He resigned from the service to become chief engineer of the DeForest Radio Company in which capacity he served until 1921. From 1921 to 1923 he practiced consulting engineering. From 1923 to date he has been associated with the Federal Telephone Manufacturing Company, since 1927 as vice-president in charge of engineering. Mr. Horle has been Chairman of the Buffalo-Niagara Section of the Institute since its inception in 1927. He was elected to Associate membership in the Institute in 1914, was transferred to the Member grade in

1923, and the Fellow grade in 1925. For several years he was a member of the Institute's Committee on Standardization. He is a member of the Engineering Committee of the Radio Manufacturers' Association.

C. M. JANSKY, JR.

C. M. Jansky, Jr., was born at Delton, Michigan. In 1917 he was graduated from the University of Wisconsin and in 1919 received the Master's degree in Physics from the same institution. Since January 1, 1920, he has been in charge of instruction in radio engineering at the University of Minnesota and of the operation of the University's broadcasting station. He was a member of the four National Radiotelegraph Conferences. He is associate professor in radio engineering at the University of Minnesota and is practicing consulting radio engineering. Professor Jansky is on the Board of Direction of the American Radio Relay League. He has contributed several papers to the PROCEEDINGS of the Institute of Radio Engineers. In 1928 he was appointed to membership on the Institute's Committee on Standardization. He was elected an Associate member of the Institute in 1918, was transferred to the Member grade in 1925, and the Fellow grade in 1928.

WILLIAM WILSON

William Wilson was born at Preston, England, on March 29, 1887. He graduated from the University of Manchester (England) in 1907 with the B.Sc. degree, in 1908 received the M.Sc., and in 1913 the D.Sc. degrees from the same institution. From 1907 to 1912 he was occupied in research work on electronic physics in the Universities of Manchester, Cambridge, and Giessen (Germany). From 1912 to 1914 Dr. Wilson was lecturer in physics at the University of Toronto. Since 1914 he has been in the Research Department of the Western Electric Company and the Bell Telephone Laboratories. During 1917-1918 he was in charge of Western Electric Company manufacture of transmitting tubes for the U. S. Government. Since 1918 he has been in charge of research, development, and manufacture of vacuum tubes, and since 1925 has also been in charge of radio research and of the development and design of the transatlantic radiotelephone equipment. He is assistant director of research of the Bell Telephone Laboratories. Dr. Wilson has served on the Institute's Committee on Meetings and Papers in 1927 and 1928. He is a member of the Executive Committee of the American Section of the U.R.S.I. and is the A.I.E.E.'s representative on the Sectional Committee on Radio of the American Engineering Standards Committee. He was elected a Member of the Institute in 1926 and was transferred to the Fellow grade in 1928.

FORTHCOMING PAPERS

The following is a list of papers on hand for publication in the PROCEEDINGS. These papers will probably be published in early forthcoming issues:

Austin Bailey, S. W. Dean, and W. T. Wintringham, "The Receiving System for Long-Wave Transatlantic Radio Telephony."

V. J. Bashenoff, Supplementary Note to "Abbreviated Method for Calculating the Inductance of Irregular Plane Polygons of Round Wire."

Robert C. Colwell, "Fading Curves Along a Meridian."

A. F. Van Dyck and E. T. Dickey, "Quantitative Methods Used in Tests of Broadcast Receiving Sets."

K. B. Eller, "On the Variation of Generated Frequency of a Triode Oscillator Due to Changes in Filament Current, Grid Voltage, Plate Voltage, or External Resistance."

A. F. Van Dyck and F. H. Engel, "Vacuum-Tube Production Tests."

J. R. Harrison, "Piezo-Electric Oscillator Circuits with Four-Electrode Tubes."

Jozef Plebanski, "Filtering Aerials and Intervalve Circuits."

E. M. Terry, "Dependence of the Frequency of a Quartz Oscillator Upon the Electrical Constants of the Circuit."

H. M. Turner, "The Constant Impedance Method for Measuring Inductance of Choke Coils."

Jonathan Zenneck, "The Importance of Radio Telegraphy in Science."

In addition to the above, the Committee on Meetings and Papers has under consideration several additional papers.

1929 BALLOTS; YEAR BOOK ADDRESSES

The ballots for 1929 officers and Board members were placed in the mails early in October. Any member who has failed to receive his ballot by this time should communicate with the Secretary of the Institute immediately requesting a duplicate.

With the ballots there were forwarded pink slips upon which members are requested to place their present business and mailing addresses. From these slips the 1929 Year Book catalog of membership will be made. It is very important that the slips be returned promptly *with the ballots.*

INCREASE IN INSTITUTE MEMBERSHIP

For the past twelve months ending with August 31st there has been a net increase in membership of 656. This creditable expansion in the number of Institute members is due to the efforts of the Committee on Membership and the individual members of the Institute. Through the increase during the past two years, the Institute has been able to do many things of direct benefit to the entire membership. The number of pages in the PROCEEDINGS during the past two years has been increased by some one hundred and twenty-five per cent. The Institute will continue to expand in all of its activities only through a continuation of the efforts of all members. If you do not have an application form to pass on to a desirable non-member, drop a line to the Secretary of your Section or the Secretary of the Institute.

Institute Meetings

TOURS TO SECTIONS

The following tours on the part of speakers have been arranged through the cooperation of the organizations with which the speakers are connected, and the Institute Headquarters office:

During September Messrs. E. T. Dickey and F. H. Engel visited the Connecticut Valley, Buffalo-Niagara, Rochester, Cleveland, Chicago, and Detroit Sections, presenting papers "Quantitative Methods Used in Tests of Broadcast Receiving Sets" and "Vacuum-Tube Production Tests."

During September Dr. Irving Wolff spoke before the Philadelphia, Washington, Atlanta, and New Orleans Sections on "Loud Speaker Characteristics and Sound Measurements."

In October Dr. Austin Bailey will speak before the Pittsburgh, Washington, and Philadelphia Sections on "The Receiving System for Long-Wave Transatlantic Radio Telephony."

R. H. Ranger will visit the Buffalo-Niagara, Cleveland, and Chicago Sections during October to talk on "Recent Advances in Photoradio."

NEW YORK MEETING

On September 5, 1928, the first regular fall meeting in New York City was held in the Engineering Societies Building, 33 West 39th Street. President Alfred N. Goldsmith presided. The 1928 Institute Medal of Honor was presented to Professor Jonathan Zenneck, of Munich, Germany, following which Professor Zenneck presented a comprehensive paper, "The Importance of Radio Telegraphy in Science."

Professor Zenneck's paper will appear in a forthcoming issue of the PROCEEDINGS.

Three hundred and twenty-five members and friends attended the meeting.

BUFFALO-NIAGARA SECTION

The 1928–1929 Officers and Committee Chairmen of the Buffalo-Niagara Section have been announced as follows: Chairman, L. C. F. Horle; Vice-Chairman, Dr. L. Grant Hector; Secretary, C. J. Porter; Chairman of the Committee on Meetings and Papers, F. Austin Lidbury; Chairman of the Committee on Membership, J. Eichman, Jr. K. Henderson has been appointed to membership on the Committee on Meetings and Papers, and J. W. Million and H. Smith are serving on the Committee on Membership.

The Buffalo-Niagara Section's Board of Direction holds regular bi-monthly meetings at which the various problems which confront the Section are discussed.

LOS ANGELES SECTION

A meeting of the Los Angeles Section was held on August 27th in the Elite Cafe at 633 Flower Street, Los Angeles. D. C. Wallace, Chairman of the Section, presided. R. H. Ranger, of the Radio Corporation of America, presented a paper on "Recent Advances in Photoradio." E. H. Hansen, of the Fox Case Movietone, read a paper on "Movietone System of Talking Movies."

Seventy members and guests attended the meeting and the dinner preceding the meeting.

The next meeting of the Section will be held on September 17th at which R. B. Parrish and K. G. Ormiston will deliver a paper, "Television Transmission and Reception," which will be accompanied by television demonstration.

PHILADELPHIA SECTION

At a meeting of the Philadelphia Section, held on September 11th in the Franklin Institute, Irving Wolff, of the Radio Corporation of America, read a paper, "Sound Measurements and Loud Speaker Characteristics."

Two hundred members and guests attended the meeting which was followed by a general discussion of the paper.

The next meeting of the Section will be held in the Franklin Institute on October 10th. Dr. Austin Bailey, of the American Telephone and Telegraph Company, will deliver a paper, "The Receiving System for Long-Wave Transatlantic Radio Telephony."

ROCHESTER SECTION

Officers and Committee Chairmen for the 1928–1929 season of the Rochester Section are announced as follows: Chairman, A. B. Chamberlain, Radio station WHAM; Vice-Chairman, Earl Karker, 143½ Plymouth Avenue South, Rochester; Secretary-Treasurer, A. L. Schoen, Kodak Park, Rochester; Chairman of the Section's Board of Direction, Ray H. Manson, Rochester; Chairman of the Committee on Meetings and Papers, Virgil M. Graham, Rochester.

SAN FRANCISCO SECTION

The San Francisco Section held a meeting on September 4th in the Engineers' Club of San Francisco. Dr. Leonard F. Fuller, Chairman of the Section, presided.

A paper by R. H. Ranger, of the Radio Corporation of America, on "Recent Advances in Photoradio" was presented by Captain Ranger. Sixty-one members and guests attended the informal dinner preceding the meeting and one hundred and thirty members attended the meeting itself.

This was a joint meeting with the San Francisco Post of the American Signal Corps Association.

The next regular meeting will be held on October 24th at which Ralph M. Heintz, of Heintz and Kaufman, will read a paper, "Radio in the Antarctic."

WASHINGTON SECTION

On September 12th a meeting of the Washington Section was held in Picardi's Cafe, 1417 New York Avenue, N. W., Washington. F. P. Guthrie, Chairman of the Section, presided. Irving Wolff, of the Radio Corporation of America, presented a paper, "Sound Measurements and Loud Speaker Characteristics."

Following the presentation of the paper, the following participated in its discussion: J. H. Dellinger, Herbert G. Dorsey, Captain Guy Hill, F. P. Guthrie, G. D. Robinson, H. H. Lyon, Marcus Hopkins, and H. J. W. Fay.

At the dinner preceding the meeting twenty-two members were present. Thirty-five members and guests attended the meeting.

The next meeting of the Section will be held on October 11 in Picardi's Cafe. Dr. Austin Bailey, of the American Telephone and Telegraph Company, will present a paper entitled "The Receiving System for Long-Wave Transatlantic Radio Telephony."

Committee Work

COMMITTEE ON SECTIONS

At a meeting of the Committee on Sections held on August 30th in the offices of the Institute there were present Donald McNicol, Chairman; E. R. Shute, E. I. Green, Quinton Adams, and F. P. Guthrie.

Correspondence covering the activities of the Sections during the summer months was discussed. At the 1929 Convention of the Institute which is to be held in Washington, D. C., during the middle of May, it is proposed to set aside a portion of the time for a meeting of the Committee on Sections with representatives from each Section present.

COMMITTEE ON ADMISSIONS

Messrs. R. A. Heising, Chairman; E. R. Shute, and H. F. Dart were present at a meeting of the Committee on Admissions held on September 5th in the offices of the Institute.

The Committee considered fifteen applications for transfer or election to the higher grades of membership.

Personal Mention

Melville Eastham, Treasurer of the Institute, returned from a three months' trip to Europe on August 20th.

Satoshi Uchida has returned to Japan, where he is located in Tokio City as engineer at broadcasting station JOAK.

S. A. Jensen, formerly of Denver, Colorado, is now in the service department of Sherman, Clay and Company of Seattle.

Since June of this year K. H. Goode has been chief chemist of the Sylvania Products Company at Emporium, Pennsylvania.

Charles E. Biele has recently been appointed assistant sales engineer of the Edison Storage Battery Company. Mr. Biele is located at Orange, N. J.

C. B. Feldman, for two years a teaching fellow at the University of Minnesota, is now in the radio research department of the Bell Telephone Laboratories, at Cliffwood, N. J.

D. K. Gifford, who has been with the Radio Corporation of America in their transoceanic telegraph department, is now at Salt Lake City, Utah, at the Department of Commerce airways station.

Carl Dreher has recently been appointed chief eningeer of R.C.A. Photophone. He remains with the National Broadcasting Company as assistant secretary of the Board of Consulting Engineers.

Dr. August Hund has been transferred from the Sound Laboratory of the Bureau of Standards to the Radio Section, in which he was formerly active for several years. He is specializing in research problems in piezo-electricity.

Bruce W. David, Secretary of the Cleveland Section, has resigned from the engineering staff of the Sterling Manufacturing Company to become connected with the Cleveland Electric Illuminating Company as assistant to the President, in charge of radio engineering.

A. M. Trogner, for a number of years in the Radio Division of the Naval Research Laboratory where he specialized in high power transmitting design and development, has left Washington to become associated with Wired Radio, Inc., of New York City.

Haraden Pratt, formerly of the Radio Division of the Bureau of Standards, is now chief engineer of the Mackay Radio and Communication Company of New York City. Mr. Pratt was engaged in aircraft radio development work at the Bureau. He has contributed several papers recently to the PROCEEDINGS on "Aircraft Radio."

Alfred Crossley, for a number of years associated with the Radio Division of the Naval Research Laboratory at Bellevue, D. C., has resigned to become chief engineer in charge of research and development of the Steinite Radio Laboratories of Atchison, Kansas. Mr. Crossley, during 1928, was Secretary-Treasurer of the Washington Section of the Institute. As Secretary, pro-tem, Thomas McL. Davis, of the Naval Research Laboratory, is carrying on the Section duties performed by Mr. Crossley.

NEW YORK-ROME AIRPLANE FLIGHT

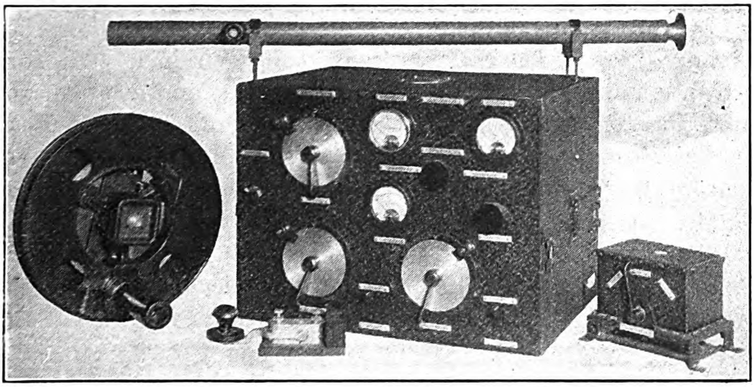
As these lines are being typed the Bellanca sesqui-plane *Roma* is poised at Old Orchard, Maine, preparatory to its attempt at a non-stop flight to Rome. The crew is composed of Commander Cesare Sabelli, pilot, Roger Q. Williams, pilot, Captain Peter Bonelli, navigator and radio operator, and Dr. Pucilli.

The radio equipment is of special design by the Radio Corporation of America. The transmitter is equipped to operate on either 600 or 45 meters. The 600-meter circuit is a Hartley oscillator inductively coupled to the antenna, and employs a UX-852 75-watt power tube which can be modulated for ICW by means

which are suspended from the fusilage by means of heavy rubber cords.

Power for the transmitter is obtained from a 500-watt generator driven by a Deslauries air fan. The generator is mounted on a retractable frame so that it may be swung into the cabin to avoid extra wind resistance to the airplane when the apparatus is not in use.

The antenna wire is of copper-clad steel, normally rolled in on an antenna reel. Two or more ten-ounce fish weights can be attached to the end of the three-hundred and fifty-foot antenna wire which is run through the fair lead shown in the photograph, on top of the transmitter cabinet.



Transmitter, Fair Lead, Antenna Reel, Flame-Proof Send-Receive Switch and Flame-Proof Transmitting Key of the *Roma*.

The receiver uses five dry-cell tubes, the circuit being a tuned radio-frequency stage preceded by an untuned coupling tube, and followed by an autodyne detector and two stages of audio-frequency amplification. The wavelength range of the receiver is from 550 to 880 meters, no provision being made to receive short waves.

Very elaborate arrangements have been made to provide for reception of signals from the plane while in flight. Several stations on the eastern coast of the United States as well as stations in England, Portugal, the Azores, and Italy will continually guard one or both of the wavelengths used in the plane.

The call letters of the plane are WRCA.

THE BYRD ANTARCTIC EXPEDITION

When the small bark *City of New York* put out from New York City on August 26th bound for New Zealand, the first unit of the Byrd Antarctic Expedition set forth for an absence of certainly one and one-half, and possibly three years. The *City of New York* was christened *Samson* many years ago. She has been plying the sealing trade off the shores of Spitzbergen for over forty years. On her voyage south through the Canal to New Zealand thirty-seven of the seventy men who will take part in Commander Byrd's antarctic adventure were present.

On September 15th the second unit of the expedition sailed from New York, when the trim ship *Eleanor Boling* (née *Chelsea*) set sail with additional men and additional equipment. Early in October the 10,000 giant *C. A. Larsen*, a veteran Ross Sea whaling steamer, leaves Hampton Roads, Virginia, with four airplanes and additional food and equipment, and at about the same time the fourth and last ship, the *Sir James Clark Ross*, another Ross Sea whaling steamer of some 10,000 tons, leaves from the West Coast with the remainder of the supplies and men.

The four ships will meet at the Ross ice barrier near the Bay of Whales (see insert map) where the *City of New York* and the *Eleanor Boling* will tie up alongside the barrier to remain as the base camp. The supply ships, *Larsen* and *Ross*, will unload and depart.

During the year following there are some four million square miles of territory, upon which no person has yet set eye, to be explored by "land" parties and airplanes.

This antarctic expedition is probably the best-equipped expedition which ever set forth; certainly the best to go to the south pole regions.

The radio equipment includes some thirty-three transmitters and receivers, with reserve supplies to keep them in operation for four or five years, if necessary.

Call letters for the various major radio units of the expedition have been assigned by the Department of Commerce as follows:

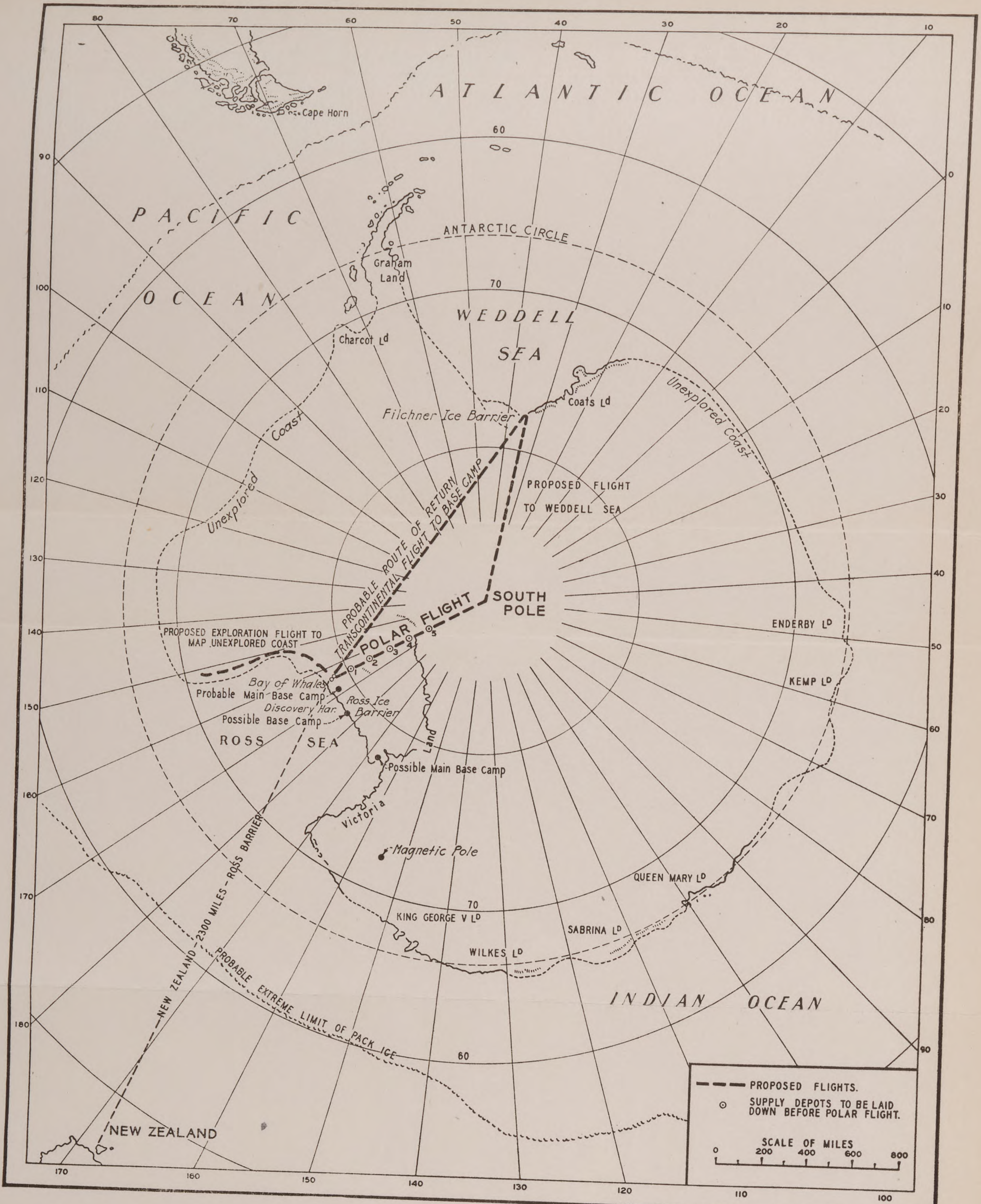
KFK—general call for any or all units of the expedition.

WFAT—the *Eleanor Boling*.

WFBT—the *City of New York*.

WFA—main base on the ice barrier (probably near Bay of Whales).

WFB—the Ford tri-motored *Floyd Bennett* airplane.

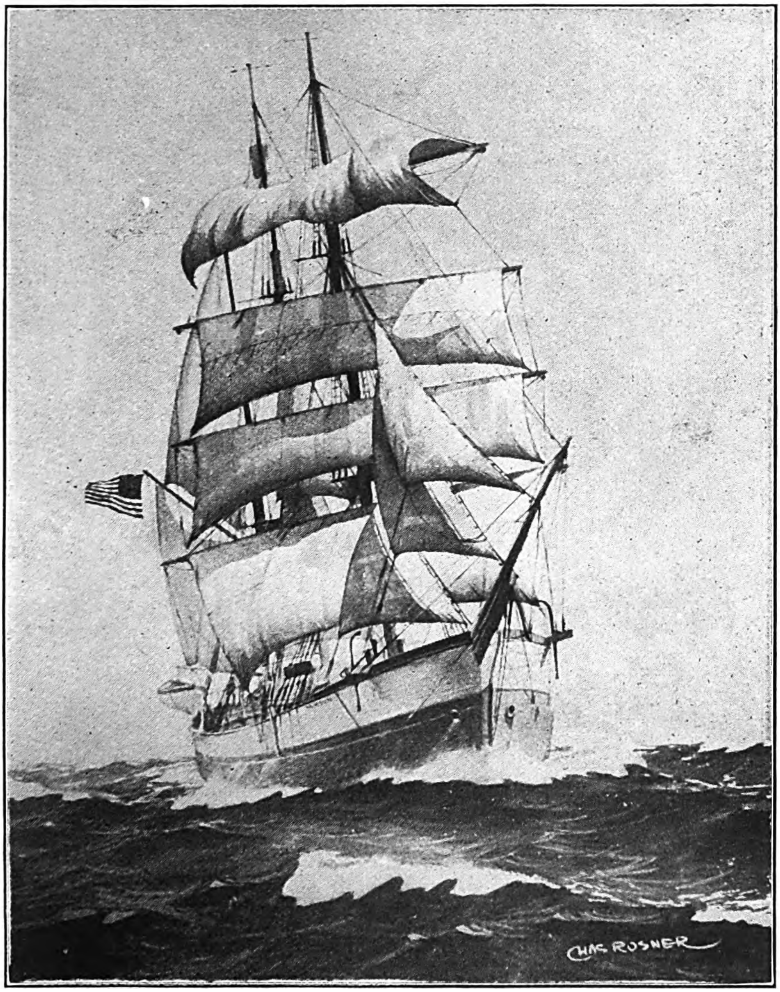


Courtesy of The Scientific American

The Antarctic Regions Showing Main Bases and Routes of Exploratory Marches and Flights of Byrd Expedition.

WFC—the Fairchild *Stars and Stripes* airplane.
WFE and WFE—advance bases on land parties.
WFF—the Fokker *Virginia* airplane.

An additional plane will use either call letters WFD or WFE.

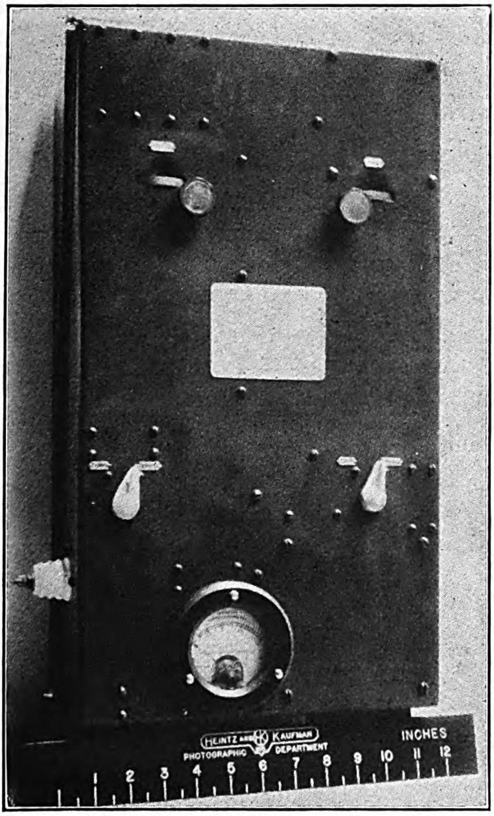


The *City of New York*, the First Byrd Ship to Sail.

The intermediate calling frequency for all units will be 500 kc and the high frequencies used for calling will be 5,600, 11,200 and 16,800 kc. The intermediate working frequencies will be 375

ten or fifteen miles from the main base camp and will be used in conjunction with the compass receiver of the other ship to fix the position of all advance parties and planes by taking simultaneous bearings from the movable transmitters.

The radio personnel includes Lieutenant M. P. Hanson, of the Naval Research Laboratory, who is chief radio engineer for the expedition; Ensign Lloyd Berkner, of the Bureau of Standards; Lloyd Grenlie, who accompanied the Byrd Arctic expedition; and Carl Petersen.

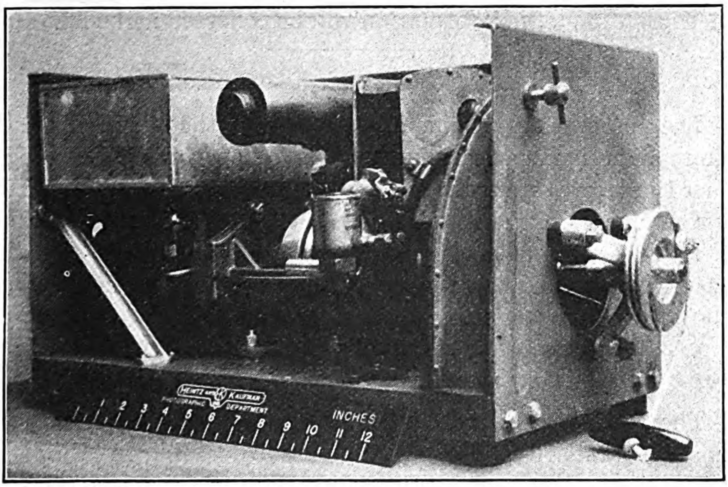


Front View of Intermediate High-Frequency Airplane Transmitter.

Both expedition ships (WFAT and WFBT) carry two transmitters, one for intermediate and one for high-frequency work. The intermediate-frequency set on the *Eleanor Boling* is an R.C.A. P-8 spark transmitter converted to a 500-watt ICW set (type

For intermediate-frequency reception (1,000 to 300 kc) standard Navy type R.F. receivers using four stages of transformer-coupled radio-frequency amplification, an autodyne detector and two stages of tuned audio-frequency amplification, are provided for both ships and for the base camp. High-frequency reception on the ships and at the base camp will be provided by the G.E.-R.C.A. high-frequency receiver using tuned r.f. amplification with the UX-222 tubes.

Each of the four planes is provided with 50-watt combined intermediate and high-frequency transmitters built by Heintz and Kaufman. Emergency battery-operated transmitters for planes and for advance parties were provided by the Burgess Battery Company. The main plane transmitters obtained plate and filament potentials from small generators driven by the plane's motor of the constant current type. The tri-motored Ford plane is equipped with a spare generator geared to a second engine. The receiving equipment for all planes uses the super-regenerative circuit. The receivers were constructed under the supervision of Hanson at the National Electric Supply Company.



Motor-Generator Unit, Driven by Single-Cylinder Two-Cycle Gasoline Engine, for Advance Parties.

A main base camp transmitter, similar to the installation on the *Eleanor Boling*, and constructed by the ships radiomen at the Staten Island shops of the Bureau of Lighthouses, will be set up near the Ross Sea.

For reception on all planes a doublet from the tip of each wing extending to the tail, with feeder wires through the fusilage, will be used. For transmission in the planes a trailing wire will be dropped from the tip of each wing.

The expedition is carrying a great deal of laboratory equipment for making fading observations, oscillograms of echo signals and much other scientific data. The fading recorder is a Westinghouse superheterodyne which was calibrated by the Bureau of Standards, and a Westinghouse oscillograph is to be used for recording echo signals.

While the main function of the radio equipment will be the provision of reliable contact with the world, it is expected that much scientific radio information will be collected and collated.

RADIO COMMISSION ENGINEERS

The permanent engineering staff of the Federal Radio Commission has been announced as follows: J. H. Dellinger, Chief Engineer; Commander T. A. M. Craven, Captain Guy Hill, F. Y. Gates, G. O. Sutton, G. C. Ross, and G. C. Blackwell.

These engineers are permanently associated with the commission and will advise it on all technical radio matters.

STANDARD FREQUENCY TRANSMISSIONS BY THE BUREAU OF STANDARDS

The Bureau of Standards announces a new schedule of radio signals of standard frequencies, for use by the public in calibrating frequency standards and transmitting and receiving apparatus. This schedule includes many of the border frequencies between services as set forth in the allocation of the International Radio Convention of Washington which goes into effect January 1, 1929. The signals are transmitted from the Bureau's station WWV, Washington, D. C. They can be heard and utilized by stations equipped for continuous-wave reception at distances up to about 500 to 1,000 miles from the transmitting station.

The transmissions are by continuous-wave radiotelegraphy. The signals have a slight modulation of high pitch which aids in their identification. A complete frequency transmission includes a "general call" and "standard frequency" signal, and "announcements." The "general call" is given at the beginning of the 8-minute period and continues for about 2 minutes. This includes a statement of the frequency. The "standard frequency signal" is a series of very long dashes with the call letter (WWV) intervening. This signal continues for about 4 minutes. The "announcements" are on the same frequency as the "standard frequency signal" just transmitted and contain a statement of the frequency. An announcement of the next frequency to be transmitted is then given. There is then a 4-minute interval while the transmitting set is adjusted for the next frequency.

Information on how to receive and utilize the signals is given in Bureau of Standards Letter Circular No. 171, which may be obtained by applying to the Bureau of Standards, Washington, D. C. Even though only a few frequency points are received, persons can obtain as complete a frequency meter calibration as desired by the method of generator harmonics, information on which is given in the letter circular. The schedule of standard frequency signals is as follows:

Radio Signal Transmissions of Standard Frequency Schedule of Frequencies in Kilocycles

Eastern Standard Time	Oct. 22[1]	Nov. 20	Dec. 20	Jan. 21	Feb. 20	Mar. 20
10:00–10:08 P.M.	550	1500	4000	125	550	1500
10:12–10:20	600	1700	4200	150	600	1700
10:24–10:32	650	2250	4400	200	650	2250
10:36–10:44	800	2750	4700	250	800	2750
10:48–10:56	1000	2850	5000	300	1000	2850
11:00–11:08	1200	3200	5500	375	1200	3200
11:12–11:20	1400	3500	5700	450	1400	3500
11:24–11:32	1500	4000	6000	550	1500	4000

[1] This schedule replaces the one previously announced for this date.

CIVIL SERVICE EXAMINATIONS FOR PHYSICIST

The United States Civil Service Commission announces open competitive examinations for positions of Associate Physicist and Assistant Physicist (salaries of $3,200 and $2,600 per year, respectively) in the following branches: heat, electricity, mechanics, optics, radio, physical metallurgy, thermodynamics and aerodynamics, or any specialized work in the field of physics not included in the foregoing.

Vacancies in the Bureau of Standards and Bureau of Mines, Department of Commerce, and under the National Advisory Committee for Aeronautics, and in positions requiring similar qualifications in other branches of the service will be filled from these examinations.

Competitors will not be required to report at any place for examination but will be rated on education and experience (70 per cent) and writings (publications, reports, or thesis to be filed with application) for 30 per cent.

Form 2600 together with further particulars contained in form No. 220 may be obtained from the U. S. Civil Service Commission, Washington, D. C.

U. R. S. I. Meeting in Brussels

The following American delegates to the 1928 General Assembly of the International Union of Scientific Radiotelegraphy, are now in Brussels, Belgium: L. W. Austin, L. M. Hull, A. G. Jensen, T. Parkinson, Maj. Gen. C. McK. Saltzman, Norman Snyder, and G. Breit.

The meetings of the General Assembly will be devoted to routine business of the organization and to the presentation of a number of scientific radio papers.

In this connection, it is announced that the volume containing the papers presented before the General Assembly of 1927, held in Washington, D. C., is now available. A table of contents of Part 1 appears as follows:

E. V. Appleton.—The existence of more than one ionized layer in the upper atmosphere, p. 2.

E. V. Appleton.—The influence of the earth's magnetic field on wireless transmission, p. 2.

L. W. Austin and I. J. Wymore.—On the influence of solar activity on radio transmission, p. 3.

Robert Bureau.—Relation entre les parasites atmospheriques et les phenomènes météorologiques, p. 6.

O. Dahl and L. A. Gebhard.—Measurements of the effective heights of the conducting layer and the disturbances of August 19, 1927, p. 16.

J. H. Dellinger.—International comparison of frequency standards, p. 18.

G. Ferrié et R. Jouaust.—L'emploi de cellules photoélectriques associées à des lampes à plusieurs électrodes, à la solution de divers problèmes concernant la mesure du temps, p. 21.

J. W. Horton and W. A. Marrison.—Precision determination of frequency, p. 28.

E. O. Hulburt.—Ionization in the upper atmosphere, p. 37.

E. B. Judson.—An automatic recorder for measuring the strength of radio signals and atmospheric disturbances, p. 38.

F. A. Kolster.—Experiences in radio compass calibration, p. 40.

H. B. Maris.—A theory of the upper atmosphere and meteors, p. 41.

René Mesny.—Note au sujet des ondes de quelques mètres, p. 42.

T. Minohara and K. Tani.—A note on the short wave long distance transmission, p. 44.

T. Nakagami and T. Ono.—Diurnal variation in signal strengths of short waves, p. 47.

T. Nakagami and T. Ono.—Seasonal variations in signal strengths of the 20 meter wave from Nauen in Japan, p. 51.

Greenleaf W. Pickard.—The relation between radio reception sunspot position and area, p. 51.

Haraden Pratt.—Apparent night variation with crossed coil radio beacons, p. 55.

George Rodwin and Theodore A. Smith.—A radio frequency oscillator for receiver investigations, p. 58.

Balth Van der Pol.—The effect of retroaction on the received signal strength, p. 63.

Robert H. Worrall and Raymond B. Owens.—The navy's primary frequency standard, p. 67.

Eitaro Yokayama and Tomozo Nakai.—The directional observations on atmospherics in Japan, p. 74.

The complete volume can be obtained for 100 French francs from the Secretariat General, U. R. S. I., 54 avenue des Arts, Brussels, Belgium.

OBITUARY

With deep regret the Institute announces the death of

Hubert Milo Freeman

Mr. Freeman was connected for a number of years with the Research Department of the Westinghouse Electric and Manufacturing Company where he conducted much original tube development work. Early in 1926 he transferred to the Radio Engineering Department of the Westinghouse Lamp Company where he continued to be closely associated with the development and engineering of receiving and transmitting tubes.

He has been an Associate member of the Institute of Radio Engineers since 1919 and has served on the Institute's committees for a number of years.

Mr. Freeman passed away at his home in Bloomfield, New Jersey, in March of this year.

ELECTRICAL PROSPECTING*

BY

J. J. JAKOSKY

(Consulting Engineer, The Radiore Company, Los Angeles, California; Research Engineer, University of Utah, Salt Lake City, Utah.)

Summary—*By means of electrical geophysical methods it is possible to obtain a great deal of valuable information regarding the extent, depth and general type of mineralization present on a property. It is not a question, however, of how much information may be obtained from a geophysical survey, but only a question of conducting the survey and examination to such a stage that gives the most optimum return on the expenditure, considering the entire development program. The experience of the writer to date is that ordinarily the most information for the least expenditure can be obtained by locating the plan view of the conductors (the mineralized zones) and determining their approximate depth and width. Such work can be done at the rate of about 8 to 25 acres per day (with the usual four men field crew) depending upon topographic conditions and amount and type of mineralization present. The information derived from such work may now be studied in connection with the geology and known ore occurrence of the district and the exploration program more efficiently carried out. In some cases, however, more complete studies were required lasting over a period of months. In addition consideration must be given the training and personnel of the company for whom the work is being conducted. A majority of the operating companies have excellent engineering and geological staffs who have a thorough knowledge of the property and are therefore capable of interpreting the results of the survey and applying it to their specific development or structural problems. In such cases the geophysicists may act only as general consultants in the interpretation of their data. At other surveys, however, it is necessary to conduct a geological examination of the property and to take an active part in planning of the exploration and development program. In every case, however, the greatest benefits are derived by proper cooperation between the geophysicists, the mining engineer, and the geologists for the interpretation of all available data. The work of one supplements and is complimentary to that of the other.*

INTRODUCTION

ENGINEERS well realize that mining is one of America's key industries, for it supplies a major portion of the raw materials used in manufacturing. Therefore, a reliable and fast means of determining whether or not an area is mineralized and warrants expenditures for exploration work should in due time have its proportionate effect in producing greater profits to the industry and resultant lowered prices to the consumer. Various types of geophysical methods have been proposed as a

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* Original Manuscript Received by the Institute, July 23, 1928. Presented before Los Angeles Section, May 21, 1928.

means for accomplishing this purpose. Magnetic, gravitational, seismic, and electrical methods have been developed for various sub-surface investigations. These methods are thoroughly scientific and based on well-known physical phenomena; they bear no relation to the old-time diving-rods or "doodle-bugs." The electrical methods are by far the most important for mining purposes judged from present application and the comparative simplicity with which the results may be interpreted. Alternating current and high-frequency phenomena play an important part in the application of the electrical methods. Such work is of interest to members of the Institute, and is opening up new fields of applications for the radio and electrical engineer.

There are several types of electrical methods some of which utilize natural earth currents, others direct or alternating applied earth currents, and still others electromagnetic induction. Of these the last has at present the widest application. Inductive methods of geophysical prospecting are capable of great versatility. They will function practically as well in swampy country as in the dry desert regions. Snow and ice are no bar to efficient work, and the character of the surface soil is unimportant, since, unlike other electrical methods, no contact with the ground or ore-body is employed. Territory covered with overburden such as lava, or float, or where no outcrops occur, can be as readily surveyed as other kinds of territory, a fact which offers immense possibilities in disclosing the location of hitherto unknown mineralized areas.

The Occurrence of Ores

The term *ore* generally denotes those *minerals* mined from the earth from which *metals* are derived. Metals seldom occur in the pure state but are usually combined with other elements or compounds, and as such are called minerals. Silver occurs combined with sulphur, and is often associated with arsenic. Copper is found combined with oxygen as an oxide, or with sulphur as a sulphide, and also as a carbonate. Lead occurs as a sulphide, as a carbonate, and as a sulphate. Iron is usually the most abundant in the earth's crust and occurs as oxides, carbonates, and sulphides. Many metals occur together and combined ores may be mined for several metals.

Ores occur in a great variety of ways or manners. The types of deposits most amenable to electrical prospecting, however, are those in which the ores fill fissures or seams in the rock and thereby form veins or ore-shoots. The mineral content of the rocks is usually deposited there by ground-water flowing through these fissures. This deposition may have been caused by cooling of the water, or by a chemical reaction between the water and the rocks, which decreased the ability of the water to hold the mineral substances in solution, causing their deposition. Such deposits may take almost any conceivable shape and may form in films, sheets, or crusts deposited upon the walls of the fissure. As a rule, the ore occurs in combination with other substances, called *gangues*, such as quartz, calcite, barite, dolomite, siderite, etc., which are very poor electrical conductors.

FIELD OF APPLICATION FOR THE ELECTRICAL METHODS

All electrical geophysical methods depend for their operation upon the effects produced by the flow of an electric current. By studying these effects it is possible to predict the general axis of current flow. The greater flow of current is in the path of greatest effective conductivity.

Electrical methods can be advantageously employed when the electrical conductivities of the different strata or components of the earth's surface differ considerably. As applied to the location of mineralized areas or zones, the greater the difference in electrical conductivity between the mineralized area and the surrounding envelope (the earth), the more pronounced or definite is the effect upon the recording or detecting instruments. Generally speaking, it is necessary that the difference in conductivity be of the order of 1 to 100 or more. Ratios of 1 to 100 are quite common; and good mineralized areas (such as massive sulphide indications) often have ratios from 1 to 1,000 to 1 to 10,000 compared with the surrounding earth.

Ores which are amenable to electrical prospecting are, generally speaking, those which may be classed as electrical conductors, though, as will be shown later, the question of conductivity is a relative one. Such ores include pyrites, graphite, chalcopyrite, arsenopyrite, some anthracite coals, a few carboniferous shales, galena, pyrolusite, magnetite, and the metals.

Many ores occur in nature in such chemical form that they are electrically non-conductive. These include the majority of the oxides, carbonates, and silicates; and two common sulphides, stibnite and sphalerite. Such ores generally are not amenable to electrical prospecting unless associated with other ores which may be conductive.

In many districts the geologist or mining engineer knows from previous investigations that certain ores often occur associated with other ores. In this way, an indirect method is available for locating the pay ore, owing to the presence of a conducting ore having no special commercial value. Unless the geology and manner of ore occurrence of a district are well-known, it is difficult to obtain an authoritative idea of the commercial values of the mineralized area. Generally speaking, electrical prospecting methods are of value only in locating mineralized areas. These usually contain ores of more than one mineral. To illustrate this point, consider a mineralized area consisting largely of zinc sulphide, a non-conductor, combined with a very small amount of iron pyrite or iron sulphide, a conductor. Owing to the presence of the pyrite, the mineralized area is amenable to electrical methods of prospecting. Unless quite close to the surface, the zinc sulphide would have no effect on the electrical readings, regardless of its mass or length. The electrical effect of such a body would depend upon the amount of pyrite present; the commercial value of the area would depend upon the zinc sulphide.

ELECTROMAGNETIC GENERATION OF AN EMF

The inductive method is so named because the current flowing in the conductor is obtained by electromagnetic induction, instead of by the use of ground electrodes through which a current is passed as in the applied potential systems, or by direct contact with the ore-body. As is well-known an alternating current flowing in a coil will create an electromagnetic field around the coil. This field will have the same frequency as the current, and will radiate or travel outward from the coil in closed magnetic or flux circuits. These flux circuits (in air or a homogeneous medium) will be perpendicular to the plane of the coil and will extend or travel outward with uniform velocity in every direction. Maximum field exists in the plane of the coil. Such a coil used for transmitting will exhibit similar figure-eight characteristics to those exhibited by the conventional direction-finding coils.

THEORY OF ELECTROMAGNETIC INDUCTION

Whenever an alternating magnetic field cuts a conductor an emf is generated. The magnitude of this generated emf is among other things a function of the strength of the magnetic field and the rate of change of the field, that is, its frequency. The higher the frequency, the greater the induced emf.

The current induced in the conductor will, of course, have the same frequency as the original current flowing in the energizing loop. Regardless of the alternating current employed—whether a so-called low frequency from 50 to 3,000 cycles per second, or the higher frequencies up to many thousands of cycles per second—practically the same fundamental phenomenon takes place.

These relationships may be expressed by the formula of electromagnetic induction:

$$E = 2\pi f M I$$

Wherein E is the induced emf.

f = the frequency of the magnetic field (or the frequency of the current flowing in the energizing circuit)

M = the mutual inductance between the primary or energizing circuit and the secondary circuit, and

I = the current flowing in the primary circuit.

MUTUAL INDUCTANCE

The mutual inductance between two electrical circuits is a constant determining the ease with which energy from the first circuit can be transferred to the second when the two are coupled by the magnetic lines of force due to the current in the first circuit. The greater the mutual inductance or coupling, the greater will be the energy transferred between the circuits. Ordinary transformer action is the most familiar example of this phenomenon. As seen from the expression given above the larger the value of M the greater is the value of the induced electromotive force.

The range in values of M for the average energizing coil and an ore-body (assuming no absorption of the field between the energizing coil and the conducting body) may be shown by considering the case of a rectangular coil and a wire of small diameter and infinite length. The coil and the wire lay in the same plane with one side of the coil parallel to the wire. If h is the length of the coil parallel with the wire, d the width of the coil, and A the distance from the wire to the near side of the coil, a direct calculation gives us for the mutual inductance M the expression:

$$M = 2h \log_e \left(1 + \frac{d}{A}\right)$$

In case of a square coil $h = d$, and $M = 2d \log_e \left(1 + \frac{d}{A}\right)$.

The magnitude of these values may best be illustrated by a practical example. Consider a vertical energizing coil 10 ft. square placed directly above the conductor so they both lie in the same plane and assuming no absorption of the magnetic field. The relationships between *M* and the distance between the coil and the wire is shown by Fig. 1. The mutual inductance decreases with an increase in distance between the coil and the wire.

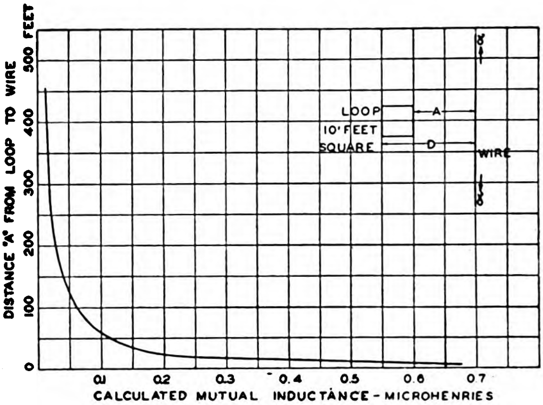

Fig. 1

When the medium between the coil and the wire is of such a character as to decrease by absorption (which is the dissipation of energy by eddy currents) the strength of the magnetic field, the value of *M* decreases in proportion to this decrease in field strength since absorption is equivalent to decreasing the coupling between circuits. The absorption in dry or desert countries may be practically neglected when operating at frequencies of 50,000 cycles per second or less to depths of 400 to 500 feet. In regions where the sub-strata are wet the absorption is quite appreciable, as will be shown in a later paragraph.

MAGNITUDE OF INDUCED VOLTAGE

The magnitude of the voltage induced in a conductor by current flowing in the coil may be calculated from the preceding data. The current flowing in the 10 ft. square energizing coil is assumed to be 5 amperes and, as before, with the coil and conductor lying in the same plane and one side of the coil parallel with the conductor. These values, for various frequencies and mutual inductance between the coil and the conductor are shown in Fig. 2. It will be noted further that the induced emf decreases rapidly with a decrease in M; that is, an increase in the distance between the energizing coil and the conductor. In geophysical applications this means that the induced emf becomes less with an increase in depth of the conductive ore-body. This is one of the factors which limits the depth at which a conductor may be located.

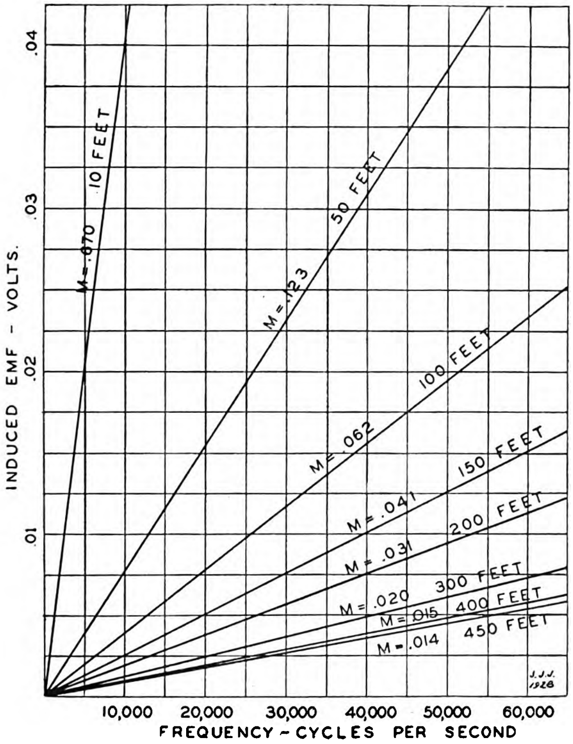

Fig. 2

FACTORS AFFECTING CURRENT FLOW

Inasmuch as electrical geophysical apparatus employing the direction-finding coil systen for the location of the underground conductive masses operates on *current flow* instead of potential difference or induced emf, it may be well to consider briefly the electrical characteristics of mineralized areas.

Mineralized zones usually possess properties quite different from ordinary solid conductors such as wires, etc. Under the practical conditions usually encountered in geophysical work, the capacitive reactance may be of very high value and in all cases where measurements or calculations have been made by the writer, the inductive component is comparatively small and completely masked by the capacitive component. This is especially pronounced in all broken, faulted, and disseminated ores, as well as in so-called massive ores containing fractures which are filled with non-conducting or very high resistance depositions such as quartz, calcite, altered feldspar or clay, and some of the oxides. These fractures may vary in size from very minute veinlets to fractures, fault zones or even igneous dikes.

The magnitude of the capacity existing between portions of a fractured ore may be seen from the following typical tests. A view of the specimen is shown in Fig. 3. This particular sample[1] is a surface outcropping on the Duprat Mines property in Kamiskotia, Canada. Samples were cut from the specimen and analyzed chemically and also submitted to a microscopic examination. The assay results, given in Table I, show that the marcasite is impure, in that it contains silicate material, lime, and alumina. The following information and the data in the table is from work being conducted at the department of Metallurgical Research, University of Utah, by R. E. Head, and R. N. Anderson, of the Radiore Company.

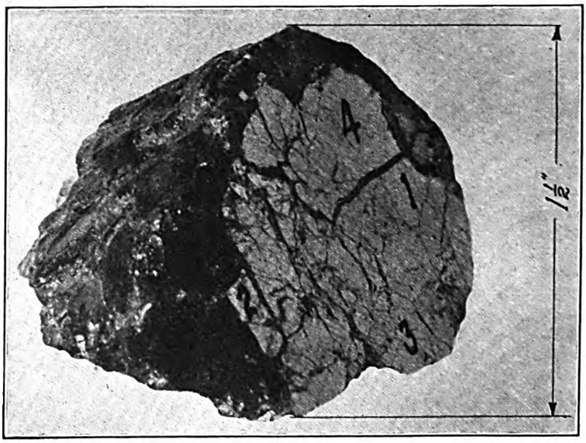

Fig. 3

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[1] Submitted by E. H. Gilford, General Manager of the Radiore Company of Canada, Ltd.

> "The microscopic examination shows that this impurity is contained in veins which fill the badly fractured marcasite. The fractures run in all directions and vary in size from very minute veinlets to 1/16 in. in width. The content of these veins is probably quartz, calcite, and altered feldspar or clay. There is also some altered marcasite in the form of iron oxide or limonite. Small specks of marcasite are also present in the veins. The marcasite is free from other sulphides and shows but little alteration, the iron oxide being only a film next to the impurities contained in the veins. Both the small and large samples were nearly enclosed in a covering of this impure material, the quartz predominating."

The capacities existing between different mineral parts of the specimen (see Fig. 3) were measured at 25 kc by means of the usual capacity substitution method.

Capacity between	Micro-microfarads
1 and 2	11 $\mu\mu$ f
1 " 3	7 "
1 " 4	7 "
2 " 3	3 "
2 " 4	7 "
3 " 4	8 "

In any circuit predominately capacitive, the impedance to current flow will decrease with an increase in frequency, with a resultant increase in current flow as may be seen by the following familiar mathematical relationship:

$$I = \frac{E}{Z} = \frac{E}{\sqrt{R^2+X^2}}$$

$$= \frac{E}{\sqrt{R^2+\left(\frac{1}{2\pi fC}\right)^2}}$$

where I = current flow
E = induced voltage
Z = impedance
R = resistance
X = reactance
F = frequency
C = capacity

It can thus be seen that as the frequency is increased, or as the capacity is increased, there is a decrease in the impedance and a resultant increase in the current flowing. This decrease in impedance at the higher frequencies more than offsets the slight increase in effective resistance due to redistribution of current caused by skin-effect. These relationships can be shown by actual measurements made on small specimens of the marcasite.

Four test specimens 1/8 in. x 1/8 in. x 1 in. were cut from samples (No. 56 and 57 from the work of Head and Anderson) and studies made of each of these specimens.

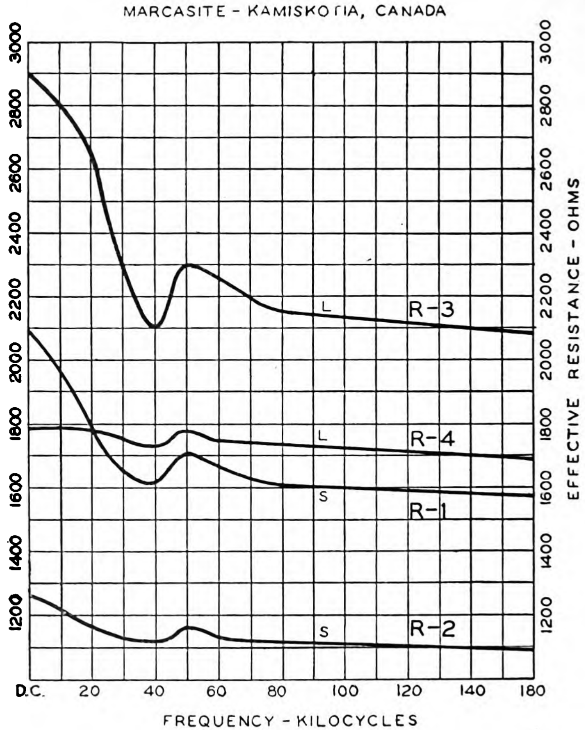

Fig. 4

Specimen No. R-1 shows a number of non-conducting veins, none of which cut entirely through it. In testing a current of 30 m-a. was initially used. This was reduced to 20 m-a. because of arcing that took place inside the specimen. The variation of impedance with frequency is shown by Curve R1 in Fig. 4.

Specimen R-2 contains less veins and consequently shows less impedance, Curve R2.

Specimen R-3 initially contained a thick transverse layer of insulating material near one end, and proved to be practically a non-conductor at the lower and intermediate frequencies. This end was cut off and constant arcing still took place within the specimen, making it difficult to obtain a balance. The results for this specimen are accurate only to about 100 ohms: Curve R3. The greater the number of fractures, the greater is the change of impedance with the change in frequency.

Specimen R4 was also practically a non-conductor due to a thick transverse insulating vein. Removing this, the sample became a fairly good conductor, Curve R4.

TABLE I
CHEMICAL ANALYSIS OF FRACTURED MARCASITE

Material	Sample No. 56 Small	Sample No. 57 Large
Iron, Fe, per cent	38.7	43.1
Lime, Ca O, per cent	0.76	0.76
Sulphur, S, per cent	42.5	44.1
Silica Dioxide, SiO_2, per cent	11.2	5.4
Aluminum Oxide, Al_2O_3, per cent	12.7	13.4
Insoluble, per cent	12.6	6.0

In observing these curves it will be noted that all of the samples of this particular marcasite had a characteristic "hump" at approximately 50 kc, and that the average impedance decreased with an increase in frequency. This decrease in impedance as the frequency is increased shows that the specimens possessed properties making them predominately capacitive.

These tests were made on a special balanced differential transformer-type alternating-current bridge designed by the author and will be described in a later paper. The results obtained on minerals and ores from various parts of the world will be given in a publication of the results of a cooperative study now being conducted by the Department of Metallurgical Research of the University of Utah and the Radiore Company.

IMPEDANCE OF DISSEMINATED ORES

Disseminated ores also exhibit quite different properties toward direct and alternating current. A disseminated ore may be considered as composed of small electrically conducting particles distributed in a matrix. As a rule this matrix is calcite, quartz, and the like, and has a low electrical conductivity; that is, a high resistance. If two conducting particles or masses are separated from one another, an electrostatic capacity exists between these two particles. This capacity is dependent upon the following factors: the geometric configuration and arrangement of the particles, and the dielectric constant or specific inductive capacity of the separating medium (in the case of a disseminated ore it would be the matrix material). A disseminated ore may therefore from an electrical point of view be considered as a resistance shunted by a number of very small capacities connected in series-multiple. Some disseminated ores have been found to have practically an infinite resistance or impedance to direct current or low-frequency alternating current, but were quite good conductors to higher frequencies of 20 kc (20,000 cycles per second) or more. Disseminated ores occurring in dry or desert regions are particularly noticeable in this respect. This is due mainly to the absence of moisture in the matrix allowing it to be practically a perfect insulator.

IMPEDANCE OF FAULTED ZONES

The action of the average broken or faulted ore-bodies when subjected to alternating magnetic fields of different frequencies is very noticeable. The impedance of such an ore-body decreases very appreciably with an increase in frequency. Such a system of conductors and effective capacities may be represented by an equivalent series parallel circuit of pure resistance and capacities, the impedance of which could be represented by:

$$Z = \frac{R}{\sqrt{1+R^2 4\pi^2 f^2 C^2}}$$

which at the higher frequencies approaches:

$$\frac{1}{2\pi f C}$$

Thus an increase in frequency causes a decrease in the effective impedance with a resultant greater current flow. In many cases such ore-bodies cannot be sufficiently energized with low frequency, but are easily energized with the higher frequencies. This may be illustrated by considering the specimen shown in Fig. 3. The estimated effective area between particles 1 and 4 is 10 centimeters. The capacity between these particles is 7 $\mu\mu$f. Assume that this same geometrical configuration is increased in size so the effective area between 1 and 4 is 25 square feet (which would be a relatively small cross-section for an ore-body). The capacity would then become 16,240 $\mu\mu f$. Calculation will show that the impedance of such a system will be 196 ohms at 50,000 cycles per second; 9800 ohms at 1000 cycles per second; and if the body were dry and the matrix non-conducting, the direct-current resistance approaches infinity.

If such a body were in a wet region its direct-current resistance would be considerably less than infinity but might not differ sufficiently from the surrounding country rock to make its detection and a study of its electrical characteristics easy.

FIELD SURROUNDING A SIMPLE CONDUCTOR IN A HOMOGENEOUS MEDIUM

An alternating current flowing in a conductor sets up an alternating electromagnetic field having the same frequency as the current. In the case of a simple conductor, such as a small-diameter wire of great length suspended alone in the air, the field will surround the wire and travel outward from it in the form of concentric circles or envelopes. The electrical geophysical methods employing direction-finding coils depend for their operation on this alternating magnetic field surrounding the conductor. The induced current flowing in the ore-body sets up a field by means of which the ore-body itself may be located.

DETECTION OF CURRENT FLOW IN A CONDUCTOR

The most satisfactory form of alternating-current detecting equipment, which allows almost direct field calculation of the location of the conductor, is that employing the direction-finding coil. By the use of such a coil the difficulties often encountered in conveniently obtaining good ground connections in the methods where equipotential curves, etc., are plotted, are overcome, especially when the ground surface is covered with heavy brush or moss, snow, and ice. Such a coil, with its associated apparatus, is easily portable and allows quick field manipulation. The complete apparatus consists of the direction-finding coil and an amplifier (for the audio-frequency range); or a detecting and amplifier set, the output of which is usually connected to a pair of head-phones. This equipment is somewhat similiar to the conventional radio direction-finding apparatus, with the special exception that it is designed for use mainly in a vertical plane (for location of conductive areas under the surface of the earth) rather than in a horizontal plane as employed in ordinary direction finding for location of stations on the surface of the earth.

"SHARPNESS" OF MAXIMA AND MINIMA

In practical direction-finding the accuracy with which a bearing can be taken depends to a large extent upon the sensitivity of the receiving apparatus to small changes in the direction of the plane of the coil. The well-known figure-eight diagram for direction-finding coils is very steep around the minimum position which means that the device will be far more sensitive as regards precision if adjusted for the zero signal instead of for the maximum signal. To this greater sensitivity of the directional properties of the coil is added the greater sensitivity of the average human ear in determining the existence or non-existence of a signal, rather than the maximum intensity of such a signal. If the readings are being made in a hard wind, or where disturbing noises are present, the operator will have difficulty in determining his minimum positions. In actual geophysical work, the minimum angles may be read accurately under average conditions to one-half degree for in-phase conditions. The question of phase relationship will be treated in a later paragraph.

For this reason the minimum point or direction is read in practice, but the direction usually recorded is that for maximum signal strength, which in a simple case would be 90 deg. from the position of minimum. The direction-finding equipment will be described later. Antenna effect and shifts in phase relationship between the field from the energizing coil (called the primary field) and the field from the conductor (called the secondary field), together with distortions of wave fronts caused by the difference in the velocity of propagation through various media must all be compensated for or corrections made to avoid errors.

PRACTICAL RELATIONSHIPS BETWEEN THE ENERGIZING AND THE DIRECTION-FINDING COIL

From what has previously been said, it can be seen that when the direction-finding coil is so placed that its axis of rotation is in the same plane and passes through the center of the energizing loop, minimum or zero signal will be obtained when the direction-finding coil is at right angles to the plane of the energizing coil. These relationships are fundamental and hold true in the elementary case, regardless of the position of the plane of the two coils. This is illustrated in Fig. 5. It is to be noticed that the axis of rotation for the direction-finding coil is horizontal *only* when the direction-finding coil is at the same elevation as the energizing coil. Initial setting up of the Radiore inductive equipment therefore consists of two steps: (1) proper alignment of the energizing coil so that its plane is always vertical and passes through the axis of rotation of the receiving coil, and (2) alignment of the direction-finding coil. The apparatus is so designed to allow this to be done accurately and quickly.

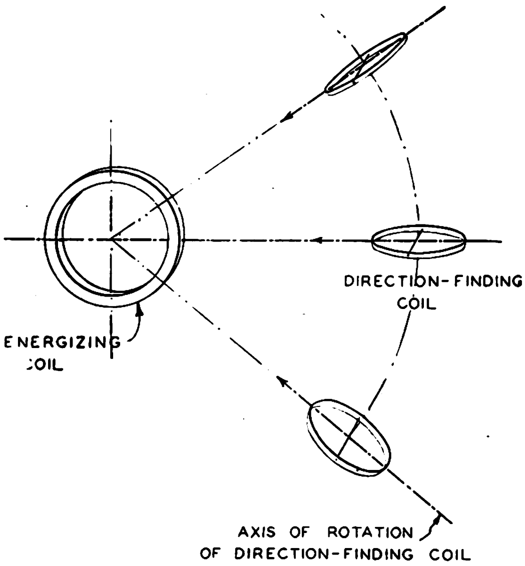

Fig. 5

When the equipment has been set up as described, the operator of the direction-finding coil knows that the position of the coil for minimum signal will normally be horizontal. Should he obtain any angle or dip other than zero (measuring from the vertical) and a strike not pointing toward the energizing coil he knows that some disturbing influence is present. This disturbing influence may be another field caused by induced current flowing in an underground conductive mass, or a distortion of the initial field from the energizing coil so it is not being propagated as would be the case were it located in a uniform homogeneous medium. These effects will be discussed later.

OPERATING CONDITIONS

When the energizing coil is placed so its field "cuts" a conductor, an emf is induced and a current flows in that conductor. As previously mentioned, this induced current sets up a field (the secondary field) of its own. A direction-finding coil placed in the vicinity of the conductor will "pick up" both of these fields.

If two electromagnetic fields of the same frequency and phase cut the direction-finding coil, the position of the coil for maximum or minimum signal strength will be determined by a single resultant of the two fields. If, for instance, one field is horizontal and the other wave is tilted so that it makes an angle of 60 degrees with the horizontal, then the direction-finding coil, under proper conditions, "points" somewhere between these two vectors the exact direction depending upon the relative strengths of the two fields. If the fields were of equal strength the resultant vector would lie equidistant between the two fields.

The elementary conditions prevailing in actual operation can best be illustrated by referring to Fig. 6. Here is pictured the plan view of a conductor of considerable length and small diameter placed so as to be in the field of the energizing coil. The direction-finding coil now has two fields linking it. At position *C* (note lower right-hand part of the figure) the component fields would exert the following effect; the energizing coil, being vertical, will tend to cause the direction-finding coil to give the loudest signal when it, too, is vertical as represented by the vertical vector. The field surrounding the conductor will tend to produce the loudest signal in the direction-finding coil at the angle shown by the vector pointing toward the conductor. The resultant effects of the primary and secondary fields are added vectorially, and the coil actually gives the loudest signal when in the position shown by the resultant vector. Moving the direction-finding coil to the position *F*, which is directly above the conductor, results in a vertical angle being obtained. At this point both the primary and the secondary fields will induce the loudest signal in the coil when it is vertical. As the coil is moved beyond the vertical position the direction of the dip angle changes, as shown by the vectors *G*, *H*, etc. Thus it can be seen that as a traverse is taken across a conductor through which an induced current is flowing, a series of dips will be obtained on the direction-finding system. For purposes of illustration, we will assume that a series of readings are being taken on a circular traverse across the surface of the ground above the conductor, as shown in the plan view. Beyond point *A*, the secondary field vectors due to the distance from the conductor are quite weak and not sufficiently strong to give a noticeable deflection to the angle of the direction-finding coil. The resultant direction for all practical purposes is vertical, or a zero dip. As the direction-finding coil is moved along the traverse toward the conductor the dip angle becomes increasingly larger until a maximum dip is reached, after which it decreases until a zero dip angle is obtained when vertically over the conductor. As the readings are continued beyond a point over the conductor the same condition results, except that the dips are in the opposite direction. This is illustrated by the arrows in the diagram entitled "Dips Across Conductor." The Dip Curve shown plots the angles to scale.

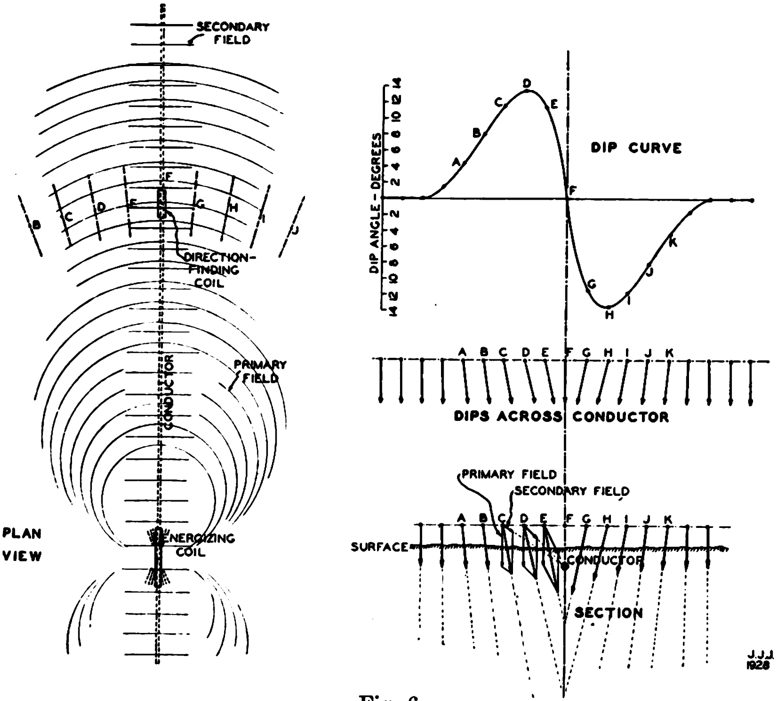

Fig. 6

In this illustration a circular traverse is shown, in order that the primary field may be constant and the change in resultant angle will be due only to variation in intensity of the secondary field and its change in angle. In practice, however, the traverses are taken along straight lines perpendicular to the conductor. The relative distance between energizing and direction-finding coils allows this to be done without any error within practical limits.

Effects of Phase Shift Between Primary and Secondary Fields

The effects of phase relationship between primary and secondary fields should be considered. If the primary and secondary fields both arrive at the receiver so that they reach their maximum or minimum values simultaneously, they are said to be in-phase. Under such conditions a definite resultant can be obtained for any given ratio of primary to secondary fields and sharp definite minimum readings will be obtained with the direction-finding loop. This can be illustrated by means of Fig. 7. The primary field is represented in magnitude and direction by the vectors *P* along the horizontal plane. The secondary field is likewise represented by vectors *S*, making, in this illustration, an angle of say 45 degrees with the horizontal plane. If the two fields are exactly in-phase their resultant will be the line *R*, and if a direction-finding coil is revolved about any axis of rotation lying

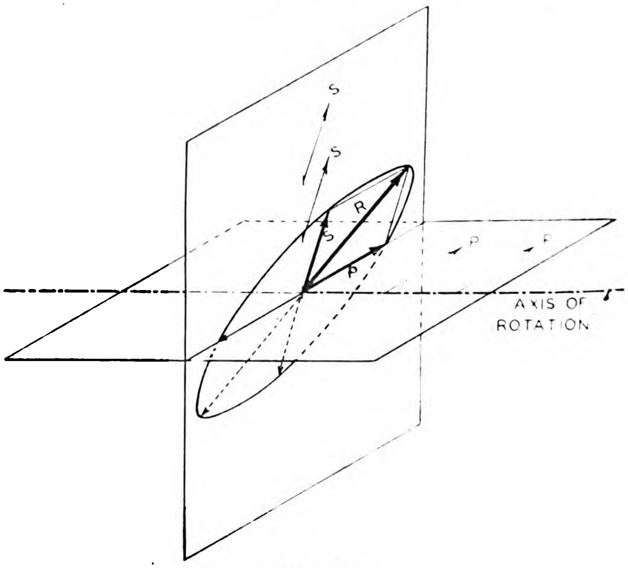

Fig. 7

in the horizontal plane and perpendicular to the vertical plane no signal would be heard in the head phones when the coil is perpendicular to line *R*.

A shift in the phase relation between primary and secondary fields is due largely to the following factors: average depth of the conductive body as compared to the distance between energizing and receiving equipment; distortion of wave front and difference in velocity of propagation between air (through which the primary field travels in reaching the receiver) and the earth (through which the energizing and the secondary field travels); transformer action whereby the current is induced in the conducting body; distribution of current in the conductor and the complex electromagnetic and electrostatic conditions existing in the vicinity of the energizing coil. When working at the higher frequencies, the phase relationship between the primary and secondary fields becomes of increasing importance. At higher frequencies the phase shift may be of sufficient magnitude to introduce a serious error in the indicated direction unless compensated for. Errors of 10 degrees or more due to phase relations may be encountered in field work[2].

Usually, however, the two fields are not in-phase and then their resultant will no longer be constant in direction and magnitude but will be of the type known as a radius-vector, the locus of which will be an ellipse lying somewhere within another ellipse passing through the extremities of the *P*, *S*, and *R* vectors. This resultant is a continually rotating vector and completes one revolution for each cycle of the alternating current. Under such a condition we find that an absolute zero signal can be obtained only when the direction-finding coil is parallel to this plane. In practice, the axis of rotation of the direction-finding coil is somewhat as shown in the figure and a zero signal cannot be obtained. Such an "out-of-phase" condition is readily recognized in practice by the operator. Out-of-phase conditions may be corrected for by numerous methods—including shifting the energizer frequency until the proper relationship is obtained at a point where readings are being made. Because of the comparatively long wavelengths used by the Radiore Company this method is by far preferable, and the energizing apparatus is so designed that proper phase relationships may be obtained merely by moving a multi-point switch.

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[2] "Fundamental Factors Underlying Electrical Methods of Geophysical Prospecting," J. J. Jakosky. *Engineering and Mining Journal*, Feb. 11 and 18, 1928.

Due to the operating characteristics of the average direction-finding equipment it is not necessary for the primary and secondary fields to be exactly in phase. It is necessary, however, that they be close enough in phase for their resultant radius-vector to be an ellipse having a major axis somewhat greater than the minor axis. The greater the difference between these two axes the sharper the reading, although an experienced operator will have no difficulty in obtaining readings of an accuracy sufficient for field operations when the ratio of major to minor axis is three or more. This ratio varies within wide limits, depending upon the strengths of the primary and secondary fields, depth of the ore-body, etc. This means that frequency of the energizer need not be shifted every time the two fields are not *exactly* in phase, but only when their resultant radius-vector does not have sufficient ellipticity for sharp readings.

If the primary field alone is present, a zero signal will be obtained when the direction-finding coil is parallel to the primary vector. Due to the fact that an infinite number of planes may be passed through a single line, inspection will show that when the primary field alone is present, the axis of rotation of the direction-finding coil may be placed in such a position as to give zero signal regardless of the position of the coil. This is one method of definitely determining the presence of a secondary field, and usually can be employed at the lower frequencies when the secondary field is too weak to allow readable dip angles being obtained.

RELATIONSHIPS BETWEEN DEPTH AND LENGTH OF ORE-BODY

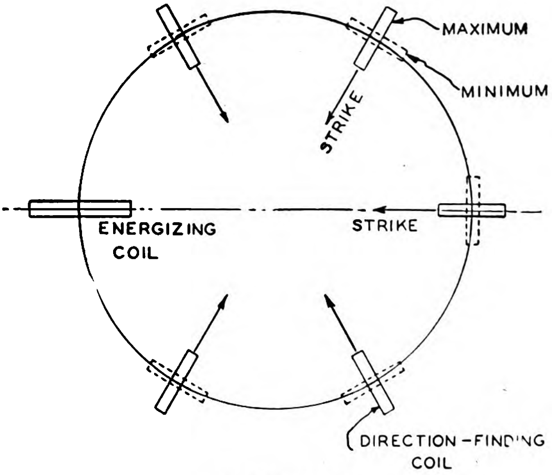
When working over a conductor of considerable length and uniform depth the ratio of primary to secondary field varies with the distance between the direction-finding and the energizing apparatus. This is due mainly to the less attenuation of the induced current when travelling over the conductor, as compared to the attenuation of the primary field. Since a certain minimum relationship exists between field strengths of the primary and secondary fields and their relative propagation angles, at which a readable dip angle may be obtained, it can be seen that the deeper a conductive ore body exists, the longer must be its effective length for optimum operating conditions. Increasing the power of the energizing system in order to impart more energy to the current induced in the conductor will not materially change conditions inasmuch as the primary field is usually increased in proportion. The greater the power used for energizing the less sensitive may be the direction-finding apparatus. The trend in design of Radiore equipment is towards lower-powered energizing apparatus and more sensitive direction-finding apparatus. This results in greater overall portability of the equipment.

The depth at which an ore body may be detected and its characteristics determined by geophysical methods depends largely upon its shape and mass. A conductive body in the shape of a sphere is one of the most difficult types to work with by inductive methods. Such a body, however, would be most optimum when using the torsion balance or other gravitational methods. A long sheet or vein conductor having the same mass as the sphere would be almost ideal for the inductive process, while its detection might not be possible with the gravitational methods, especially if it existed at any depth.

LENGTH OF ORE-BODIES

The *electrical* length of a conductive ore-body varies with the type of mineralization and the frequency of the energizing current. If the body is a massive mineralized area containing no faulted zones or fractures filled with an electrically non-conducting material, the body will be electrically conductive to high- or low-frequency current and direct current. If, however, the body is fractured, broken, or faulted and replacements or depositions of calcite, lime, quartz, etc., have taken place in such a manner as to insulate various portions of the body, it will behave quite different electrically with currents of different frequencies. When using high frequencies the induced current will readily pass through these breaks due to the electrostatic capacity existing between the various parts of the body, as has been discussed in an early part of this paper. Low-frequency currents will suffer much greater impedance and considerable difficulty will be obtained in energizing such a broken body. It will thus be seen that the effective length of an ore body may vary with the frequency. The change in energizing frequency from low (1000 cycles) to high (50,000 cycles or more) will furnish considerable information as to the structure of the body and has been successfully applied in any number of cases.

It must not be assumed, however, that breaks or faults are not detected by using the high-frequency equipment. Such breaks are clearly indicated by reversal and shape of the index curve due to the change in ratio of primary and secondary fields, and the shift in phase relationship at the break or faulted zone. The phase relationship existing between various parts of a broken conducting body depend upon their relative length, the mutual inductance between those component bodies and the energizing system, the effective resistance of the separating medium and the operating frequency. At best, it is a complicated relationship. Detailed studies have been made of numerous faulted or broken zones and absolute check obtained between the electrical readings and the geologic conditions.



PLAN VIEW

Fig. 8

Under such conditions the high-frequency equipment possesses another advantage in that one set-up of the energizer will usually be sufficient to read the entire body, while when using the low-frequency source of power it is usually necessary to move the energizing apparatus past each broken or faulted zone. This is of especial disadvantage when the conducting bodies are of short length as compared to depth, as difficulty is usually met with in detecting the presence of a secondary field in the region of the energizing coil where a strong primary field exists.

STRIKE READINGS

The azimuth angle or direction of the direction-finding coil when in vertical plane, is called the "strike," and represents the resultant direction of the fields cutting the direction-finding coil with reference to a horizontal plane or plan projection. The relationships between the relative position of energizing coil and the direction-finding coil may be seen by referring to Fig. 8. It will be noticed that the strike of the direction-finding coil is *toward* the energizing coil *only* when it lies in the plane of the energizing coil. When the direction-finding coil is not in the plane of the energizing coil, the directions for strike are as indicated. This is one of the reasons why the energizer in Radiore practice is always "pointed" toward the direction-finding apparatus. When in this position, the receiving operator knows that any strike other than toward the energizing equipment is caused by some distorting influence which may be due to distortion of the primary field wave-front, as will be described in more detail later, or a secondary field created by an induced current flowing in a conductor underground.

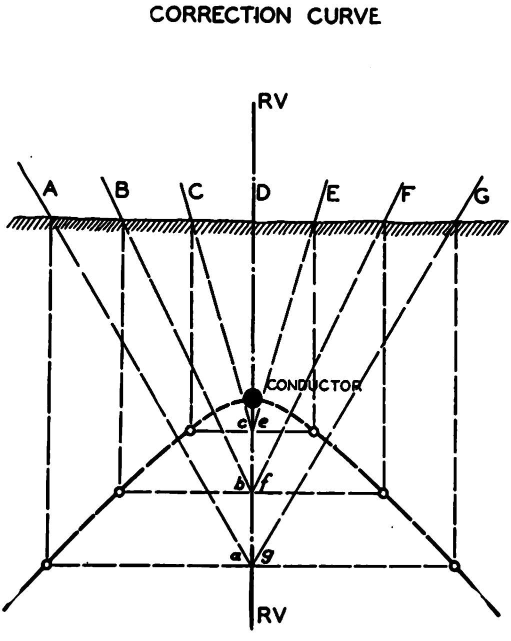


Fig. 9

When the energizer is placed in the vicinity of a conductive zone, the secondary field induced in such a zone will have a certain direction with reference to the primary field and cause a distortion in the apparent strike from where it would be normally, were that field not present. Strike and dip angles are two of the important electrical phenomena to be noted in a reconnaissance survey for the indication and resultant location of a conductive mass or zone.

DETERMINING DEPTH OF CONDUCTOR BY CURVE

A previous paper[2] described an empirical method of plotting to obtain the depth of a conductor. Briefly this consists essentially of the procedure illustrated in Fig. 9. Through each observation point on a given traverse lines are drawn making an angle with the vertical equal to the dip angle. Through the points of intersection of these lines with the vertical *RV* at which zero dips were obtained at the surface, lines are drawn parallel to the surface. Then perpendiculars are dropped from each observation point to its corresponding horizontal line. A smooth curve drawn through the intersections of these horizontals and verticals is the so-called correction curve. The apex of this curve is the position of the conductor.

Plotting the correction curve as described virtually amounts to assuming that the secondary field vector is horizontal and that the primary vector is vertical and that a definite relationship exists between the magnitude of these two vectors. In this position the secondary vector is between the upward angle produced by distortion of wave front (to be described later) and the downward angle caused by drawing a smooth curve which neglects the "tip" that normally belongs on the complete curve of the general type: $y = Ke^{x} + K'e^{x}$. The relative magnitudes of the primary and secondary field vectors is taken into account in actual practice by proper distance of operation between the energizing and direction-finding apparatus. This distance is dependent upon the energizing power used, approximate depth of ore-body, conductivity of overburden, and type of mineralization. This type of curve need not be considered as empirical, however, in location of the plan view of the conductor. The reversal or change in direction of the dips at a point immediately over the electrical axis of the conductor is indicated by the point of inflection of the curve, and is usually further checked by plotting the dip curves, etc. The curves are therefore an ex-

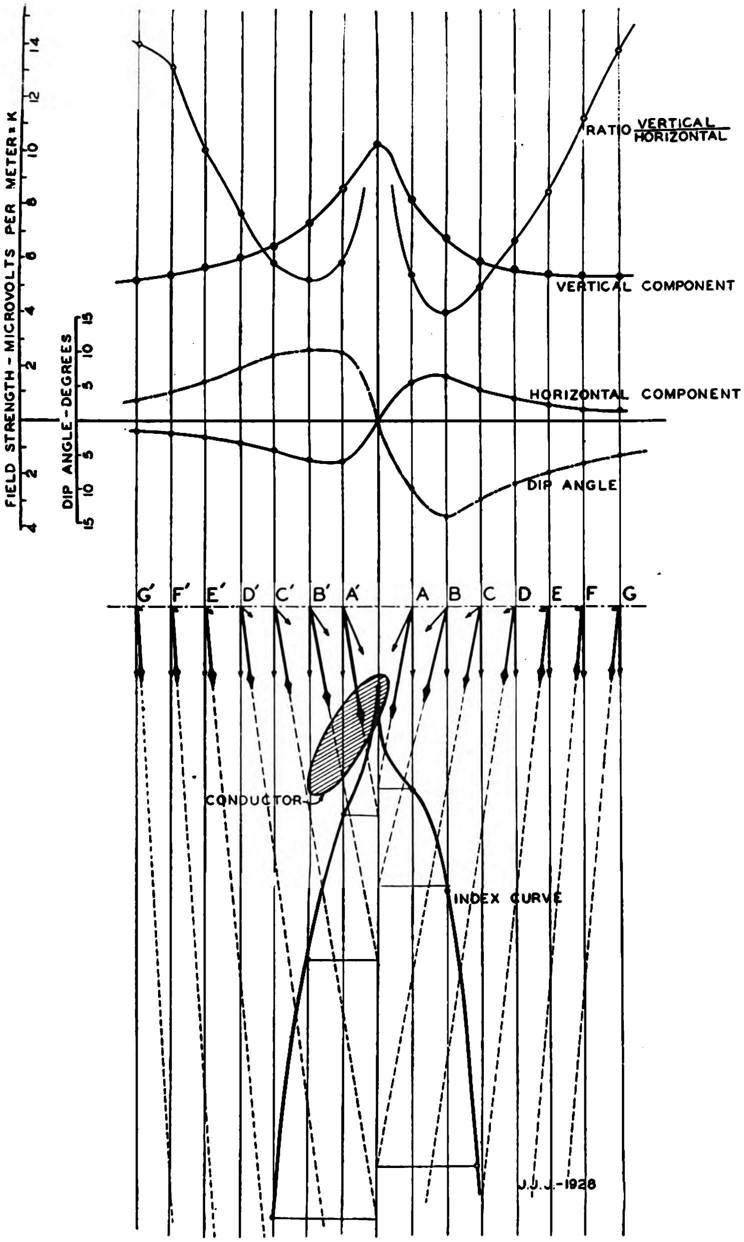

FIELD STRENGTH - MICROVOLTS PER METER = K
14
12
10
8
6
4
2
2
4
DIP ANGLE - DEGREES
15
10
5
5
10
15
RATIO VERTICAL / HORIZONTAL
VERTICAL COMPONENT
HORIZONTAL COMPONENT
DIP ANGLE
G′
F′
E′
D′
C′
B′
A′
A
B
C
D
E
F
G
CONDUCTOR
INDEX CURVE
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Fig. 10

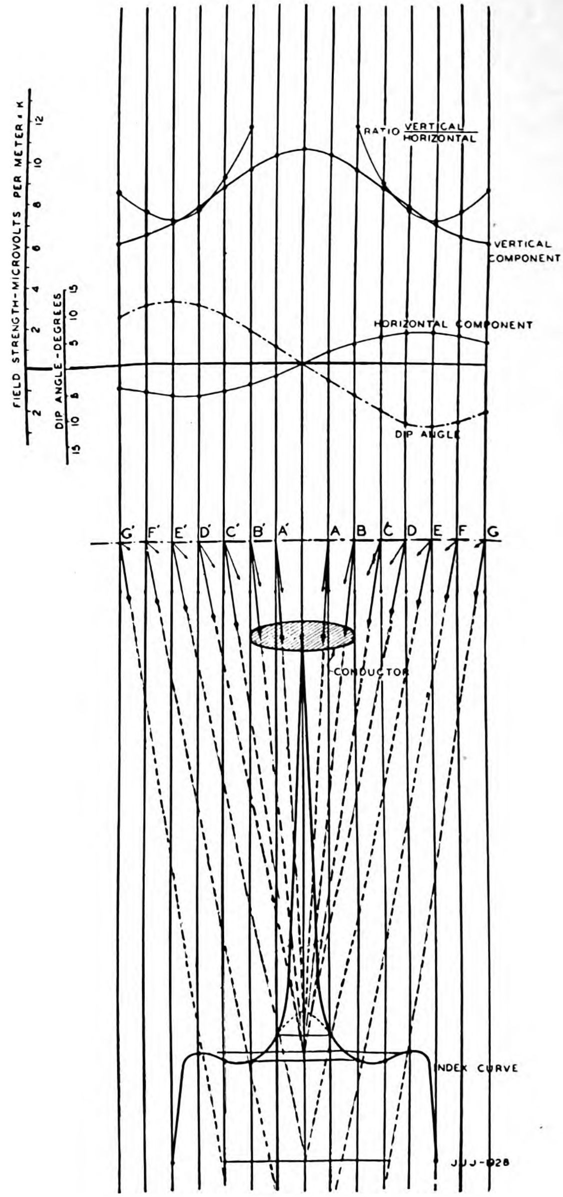
FIELD STRENGTH-MICROVOLTS PER METER × K
2
2
4
6
8
10
12
DIP ANGLE-DEGREES
15
10
5
5
10
15
RATIO VERTICAL/HORIZONTAL
VERTICAL COMPONENT
HORIZONTAL COMPONENT
DIP ANGLE
G'
F'
E'
D'
C'
B'
A'
A
B
C
D
E
F
G
CONDUCTOR
INDEX CURVE
JJJ-1928

Fig. 11

cellent graphic method for determining the plan view of the conductor, and allow the field operator to determine something of the nature of his conductor by the general shape of the curve, as well as to check any errors in field readings. By ascertaining the depth of the electrical axis of the conductor at different traverses, it is possible to draw a profile view of the entire conductor.

STUDY OF ELECTRICAL READINGS

Various methods may be used in studying the data obtained from an electrical survey. Graphical methods are more rapid and by their use considerable information can be gained of the nature and geometrical configuration of the conductive body. Curves characteristic of certain types of ore bodies have been analyzed mathematically and completely verified by field measurements and subsequent mining operations. The applications of such curves will be illustrated in a general way for three cases of using an elliptically-shaped conductive body, assuming even current distribution, no adjacent conductors, or wave distortion.

In Fig. 10 is shown the conductor "on edge," and traverse readings made at stations, *A*, *B*, *C*, *A'*, *B'*, *C'*, etc. The index curve for this type of conductor is as shown. Above this are plotted the dip angles, the vertical and horizontal component of the vectors and the ratio of vertical to horizontal components. Note the general shape and symmetry of the curves.

In Fig. 11 is shown the same elliptical ore body having its major axis horizontal and with its geometrical center at the same distance below the surface as in the preceding illustration. A decided difference can be noted in the shape of the index curve, as well as the values of the other components.

In Fig. 12 is shown the same shape conductive body making an angle of approximately 30 degrees with the vertical. The unsymmetrical curves immediately show this condition, and also indicate the general direction of the dip of the ore body. Such a conductor will present an unsymmetrical electrical arrangement when readings are made at the surface of the ground and the strike and dip curves will be unsymmetrical in appearance. By proper interpretation of dip and strike curves it is possible to determine the angle of dip of a mineralized zone. Especially is this valuable in regions where overburden exists and the structural geology of the underground structure is not definitely

known. In such instances it is quite impossible to plan the proper diamond drilling program without knowing the dip or at least the direction of the dip of a conductor. Inspection of the dip curve shown in the figure will indicate that the diamond drilling should be on the side of the conductor having the smallest dip angles. Actual field practice has proven that these theoretical curves check very well with the results of field practice, and that after drawing the index curves it is possible to predict the general shape and depth of the ore-body, well within reasonable mining tolerances.

FIELD PROCEDURE DURING DETAILED SURVEY

After the electrical conducting or mineralized areas of a property have been determined in the reconnaissance survey, a detailed survey is usually conducted to ascertain the depth of the electrical axis and the characteristics of the conducting zones. The energizing apparatus is set up over the conductor and the plane of the coil is placed in the general direction or strike of the conductor. Direction-finding coil readings are then made at various stations called *traverses* along the conducting zone or "indication," by the procedure described in connection with Figs. 6 and 9. Eight or ten readings are taken with the direction-finding coil at each traverse. The traverses are from 40 to 200 ft. apart, depending upon local conditions and the information desired. During each set-up of the direction-finding and energizing apparatus the operators must have the energizing coil vertical, the plane of the energizing coil passing through the center of the direction-finding coil, and the axis of rotation of the direction-finding coil passing through the axis of the energizing coil. Resultant strike and dip angles, distances between energizer and direction-finding coil, traverse and station numbers, topography, etc., are recorded. For certain work the power and operating frequencies are recorded in addition.

COMPLETED SURVEY MAP

A portion of a map of a typical electrical survey is illustrated in Fig. 13, showing data from a survey conducted in Inyo County, California. The "indication" in the plan view is tied in by the usual surveying methods to known property corners, bench marks, and other features, to allow later location of the indication should the electrical survey stakes be removed. In drawing the profile view of the conductor the depth of the conductor is obtained by plotting correction curves for traverses taken at intervals along the *indication*. By drawing a curve through the indicated depths at each of these traverses, the electrical axis of the conductor is located. The correction curves

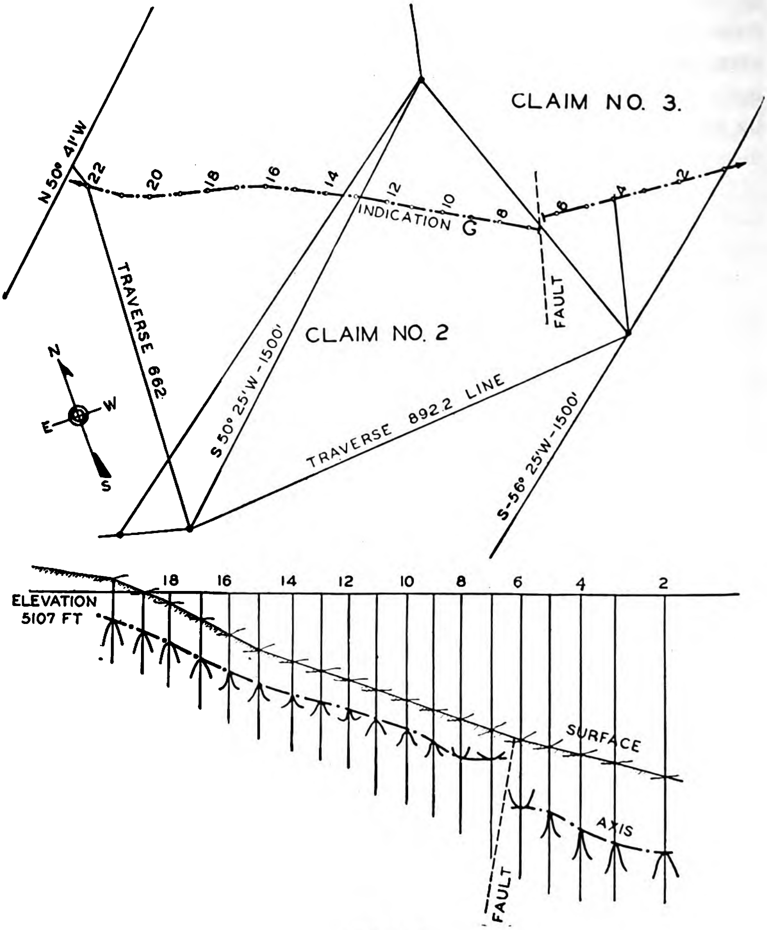

Fig. 13

for each traverse are shown in the figure. These curves as shown are drawn by imagining the traverses as being rotated 90 degrees to allow the curves to be plotted in the plane of the paper. Note the reversal of the index curves at the fault, and the displacement of the conductor.

DISTORTION OF SECONDARY FIELD

In an ideal case the magnetic field surrounds the conductor and travels outward from it uniformly in all directions. Under such conditions the direction-finding coil can be used to locate by triangulation the position of the conductor where two or more different readings have been made. In actual practice, however, the wave front traveling outward from the conductive ore body is not a true circle, and the conductor cannot be located by the simple process previously described. The wave front is distorted because of the existence of three factors: irregular shape of the conductor, irregular current distribution in the conductor, and difference in velocity of propagation for the wave in penetrating different media.

As is well-known, the velocity of propagation of an electromagnetic wave varies with the character of the media through which the wave passes. Measurements made in various locations give values of the velocity of propagation in air to that through the earth from 1.5 to 4.5. In other words, the wave travels from 1.5 to 4.5 times as fast through the air as through the earth.

In Fig. 14 is plotted the wave front for conditions where a 4 to 1 velocity ratio exists. Assuming an ideal condition where the earth is homogeneous, with the secondary field alone present and the wave fronts as indicated, the position of the direction-finding coil for maximum signal strength will vary with the distance from the vertical position above the conductor; hence, it will be noted that the curve is not a circle after the wave emerges above the effective surface of the earth. When immediately above the ore-body the direction of maximum signal strength is downward, and toward the ore-body. A reading made at point *B* would give an intersection at point *b*. A reading at *F* will give an intersection at *f*, which is above the surface of the ground. It must be remembered, however, that the intersections obtained in practice when the conventional vertical energizing coil system is used are due to the resultant effects of the primary and secondary fields. The presence of the primary field usually causes the resultant intersections to be below the surface of the ground.

It will be noted that distortion of wave front tends to give an indicated depth less than that of the actual conductor. This factor has been considered in the working out of the empirical *correction curve* already described.

the height of the energizing coil above the surface of the earth. The distortion increases, however, with the frequency, but within the range of operating frequencies and the range in earth resistivities of the overburden encountered in practical geophysical applications, the increased distortion due to the employment of the higher frequencies has been found to be well within experimental error. The question of frequency will be discussed later.

PHANTOM DIPS

Owing to the distortion of the primary field or improper alignment of energizing and receiving equipment, it often happens that a small (usually less than 20 deg.) "dip" or improper strike direction is obtained. These are called phantom dips or strikes and are readily recognized by the experienced operator. Such dips and strikes are often obtained when the energizer and direction-finding coils are located on a ridge, in a narrow valley or canyon, or at the edge of a deep cut or precipice. Usually the greater the distance between the energizer and the direction-finding coils, the greater is the wave-front distortion. This accounts for the Radiore practice of keeping a relatively short distance between energizing and direction-finding apparatus. Such a system, of course, requires many more "moves" or "set-ups" of the energizing equipment per day, but this was considered in making the energizing equipment quite portable and quick to set up.

Referring to Fig. 15 we have an illustration of distortion of a wave front due only to the fact that the energizer was placed "back" and a little to one side of a conductive hill. The heavy dot-and-dash line shows the position of the actual wave front. No secondary field is present. Note that the "strike" of the direction coil does not indicate the true direction of the energizer. The right-hand view (which looks from point *C* toward *O*) shows that the wave front is distorted vertically as well as horizontally. At that point a vertical distortion or phantom dip of 18 degrees is obtained, and a strike error of 27 degrees.

EFFECTS OF OVERBURDEN

At this point the question of relative conductivities and directional effects may be considered. In desert countries the highly resistant nature of the overburden makes the secondary field from the induced current flowing in the overburden negligible. Even in wet regions where the ratio of conductivity of ore to overburden may be only of the order of 100 to 1, the secondary field of the overburden will not mask that of the ore because of the large difference in conductivities. The currents flowing in the ore-body are so large compared to those in the overburden that the field due to the latter practically never masks that due to the former. It will readily be apparent that the field caused by the current in the ore-body will have a definite or well-defined axis while that from the currents in the overburden will not have such an axis. The result is definite indications of maxima and minima on the proper type of direction-finding apparatus for the ore-body field and no definite converging dips or "indications" for the overburden field. The field from the overburden tends of course to minimize that from the ore-body, and proper corrections must be applied for the depth calculations.

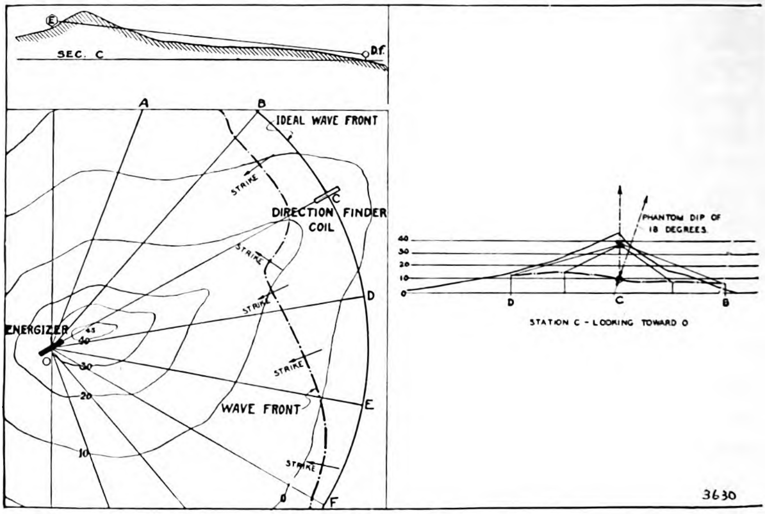

Fig. 15

EFFECTS OF FREQUENCY

The high-frequency geophysical methods described in this paper differ from the conventional inductive methods in that a comparatively high frequency is used, i.e., 30–50 kc. These frequencies were adopted after a series of tests to determine the best frequencies for the complex electrical and physical conditions met with in field practice. The various electrical geophysical methods depend for their success on the existence of a difference in electrical conductivity between the ore-body and the overburden. Granting that there is a difference in conductivity the problem is to cause a current to flow through the ore-body. Other things being equal a higher frequency means a higher induced emf not considering the decrease in magnetic intensity due to absorption of the magnetic field by the overburden. If the conductivity of the overburden is appreciable the question is whether the increase in the emf at a high frequency is more than compensated for by the decrease due to absorption. It must be remembered that absorption takes place at all frequencies both high and low. There is a balance between an increase in emf due to an increase in frequency as compared to a decrease in field strength with the increase in frequency.

Without going into a detailed mathematical discussion, it is possible to illustrate the effects of high frequency by means of a numerical example. Applying a theoretical absorption equation[4]:

$$H = H_0 e^{-2\sqrt{\frac{f}{\sigma}}Z}$$

where H is the magnetic intensity at any depth Z; H_O is the intensity at the surface, f is the frequency and σ is the specific resistance of the overburden. We find that there is an increase in the induced emf at the higher frequencies. For purposes of illustration consider an overburden having the comparatively low specific resistance of 10,000 ohms per centimeter[3] and for a depth of 100 feet (3048 cm). If we assume first a frequency of 10,000 cycles we find the intensity of the field at that depth is 52 per cent that at the surface. If we increase the frequency five-fold (to 50,000 cycles) we find that the intensity is 26 percent that at the surface. Thus increasing the frequency five times only doubles the absorption. If we introduce these facts into our equation $e = 2\pi MfI$ we find that e has increased 2.5 times. Now if we consider that the high frequency will also be

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[4] Also see "Principles of Radio Communication," by John H. Morecroft, John Wiley and Sons, 1927, pages 146 and 844.

"Electricity and Magnetism," by J. H. Jeans, Chapter 15, pages 473-479; Cambridge University Press.

"Transient Electric Phenomena and Oscillations," by C. P. Steinmetz, Chapter 6, pages 361-374; McGraw-Hill Publishing Co.

apparatus, operating at frequencies from 30 kc (30,000 cycles per second) to 50 kc. The wire or coil through which the high-frequency current flows is enclosed within the waterproof housing or "doughnut" *A*. The lower part of the loop case contains the frequency changing apparatus shown in the insert. This apparatus consists essentially of a high-frequency oscillator using two 7½-watt tri-electrode vacuum tubes *B* operating on each half of the 500-cycle supply. Proper plate and filament potentials are obtained from a transformer *C*. The entire equipment, including tube sockets, transformer, blocking and loading condensers *D*, grid resistor *E*, radio-frequency ammeter *F*, leveling bubbles *G*, frequency changing switch *H*, and marine connection plug *I*, etc., is mounted on an aluminum chassis or cover-plate *J* as shown. A glass-covered waterproof "port-hole" *K* is provided for observing operation or changing of the tubes.

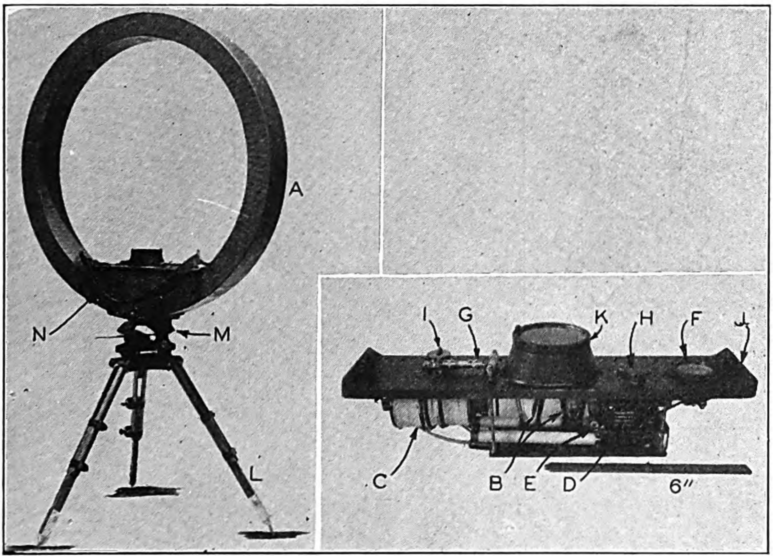

Fig. 17

The energizing apparatus is set on a tripod *L* having a graduated turntable and universal leveling attachment *M*.

In moving the apparatus from place to place a shoulder strap *N* (also see Fig. 18) is provided, and the tripod swings up alongside the coil. This allows the apparatus to be quite portable and minimizes the time required to make readings. The entire energizing apparatus weighs about 55 pounds and may be readily carried and handled by one man.

Fig. 19 shows a low-frequency coil. This coil consisting of a number of turns of rubber-insulated stranded wire is bound together into a single cable to facilitate handling. The flag shown is placed at the electrical center of the coil and is used for proper alignment of the direction-finding coils.

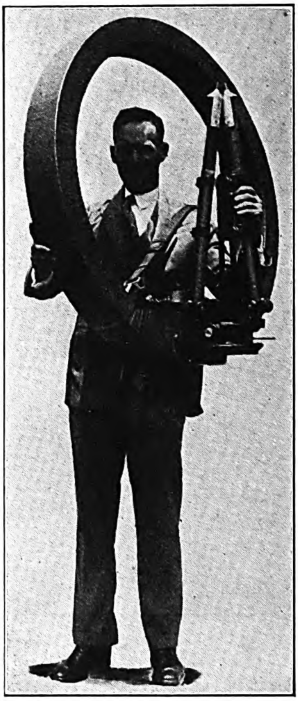

Fig. 18

Power supply for both the high- and low-frequency systems is obtained from a specially designed 600-cycle 150-watt hand-cranked alternator, a disassembled view of which is shown in Fig. 20. Two handles for cranking are provided, and connected to a sprocket chain and gear drive on the alternator. The alternator turns over at a speed of 4000 r.p.m. Ball bearings are provided throughout and the outfit cranks easily for the generation of the necessary power. Approximately 50 watts is required for the high-frequency apparatus and 150 watts for the low-frequency loop. Field excitation is supplied by three

No. 6 dry cells in series with an automatic switching arrangement which closes the circuit while the cranks are being turned. A flywheel is placed on each side of the rotor of the alternator, the whole rotating assembly being enclosed in an aluminum dust-proof housing. A waterproof steel case, provided with carrying straps and tump-line, encloses the entire unit. This outfit, complete with batteries, connection cords, a.c. and d.c. indicating meters, etc., weighs approximately 60 pounds.

In actual field operation the outfit of course is cranked only while a reading is being made with the direction-finding coil. About 5 seconds time is required to make a reading by an average experienced operator.

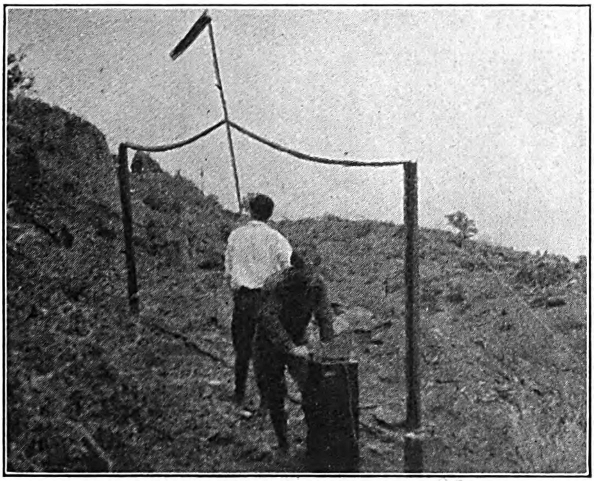

Fig. 19

DIRECTION-FINDING APPARATUS

The most recent direction-finding apparatus developed is illustrated in Fig. 21 showing the high-frequency and low-frequency equipment. The mounting head *A* on which the direction-finding coils are mounted is provided with a slighting arrangement *B* (similar to gun peepsights), whereby the axis of rotation *C* of the coil may be aligned quickly with the center of the energizing coil. A graduated vertical arc *D* is attached to the pivoted plate *E* holding the direction-finding coil. The vertical angle index mark and level bubble arc are attached to the movable arm adjusted by thumb screw.

The head is rotatable on a vertical axis and azimuth angles may be read on a scale *H*. The entire assembly is mounted on a ball and socket plate to permit rapid set-up in rough territory.

The same mounting head and set-box are used for both low- and high-frequency work; it is only necessary to fasten the proper direction-finding coil on the rotatable plate with two thumb screws. A 600-cycle coil is shown in the center view of Fig. 21, and a 30- to 50-kc coil in the righ-hand view.

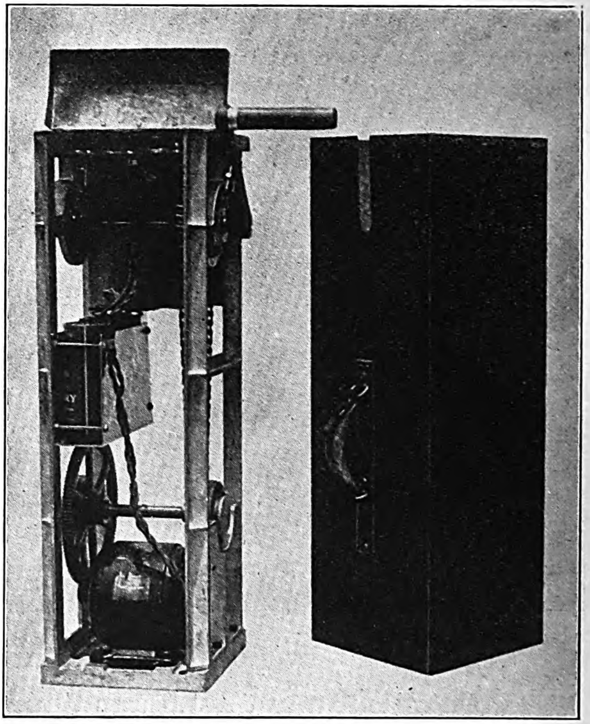

Fig. 20

The direction-finding coil is electrically connected to the vacuum-tube set box containing a detector (so arranged that it functions as an audio-frequency amplifying stage when using the low-frequency coils) and a two-stage peaked transformer-coupled audio-frequency amplifier. The entire apparatus, (including the heterodyne control dial, compensating controls, and batteries) is contained in a waterproof aluminum box $4\frac{1}{2}$ in. × $4\frac{1}{2}$ in × 17 in. (See Fig. 22.) A double range voltmeter is placed

A view of the complete apparatus is shown in Fig. 23. Two direction-finding outfits are usually worked with each energizer. The energizer operator alternately aligns his coil while the reading is being made with first one outfit and then the other.

FIELD OPERATING CONDITIONS

The design of the equipment is such as to allow maximum flexibility, and to be of efficient use under the various climatic conditions encountered in field practice.

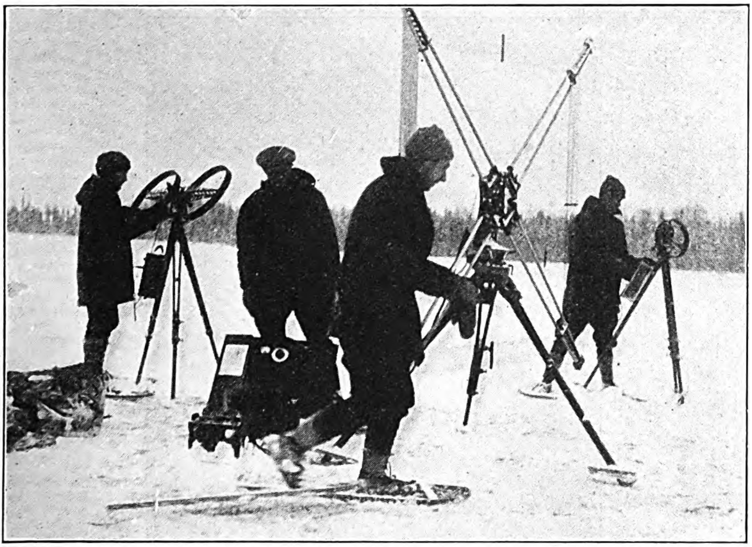

Fig. 25

In Fig. 24 is shown one type of equipment in use on the Minnietta Property of the Modoc Mining District in California. This is within a few miles of Death Valley and is quite a rugged country. The other climatic extreme is encountered in the work in the province of Quebec, Canada. The field crews operate through the winter and snowshoes are used for both men and apparatus tripods. The tripod shoes consist of cast aluminum plates 6 in.×8 in., fastened to the tripod leg about 6 in. above the tip, Fig. 25.

Operation in swampy country or lakes is not uncommon as many mineralized areas exist in such territories. Fig. 26 was taken during a recent survey over a swamp in Southern Canada.

PERSONNEL OF CREWS

The field crews usually consist of four men; a mining engineer, a geologist, an electrical engineer or trained radio technician, and a field assistant. As a rule all of these men are technical graduates.

INTERPRETATION OF FIELD DATA

One of the most important steps in geophysical prospecting is the proper interpretation of the field notes. The fundamentals of geophysical prospecting are well-known and all relationships for ideal conditions may be theoretically expressed by mathematical formulas. The proper interpretation of data, however, contains many empirical steps or operations and can only be done in the light of experience. At first glance it may seem inconsistent that the interpretation of data contains many empirical steps when the fundamentals are definitely known. This is due to the fact that conditions existing underground are quite complex. Take for instance only one of the many factors to be considered; the occurrence of ore. If conductive orebodies occurred in uniform diameter straight cylinders and in

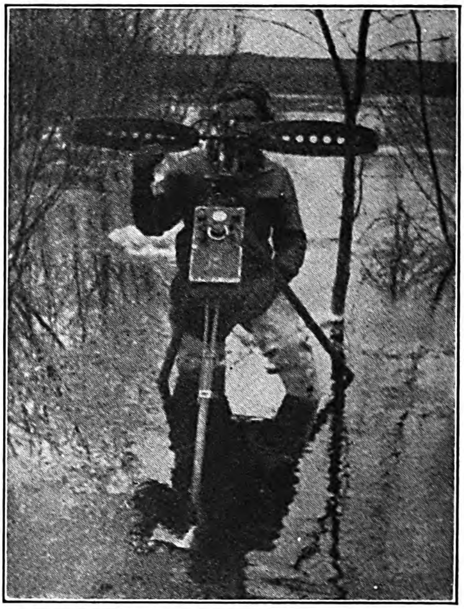

Fig. 26

a homogeneous medium, the problems of depth, etc., would be relatively simple. Actually, however, mineralization occurs in anything but uniform shapes. A mineralized zone may be quite large in one place (ore shoot) and then diminish to a very small cross-section (stringer, pinch-out) at another place. Numerous connecting zones (fingers, shoots) branch out from the main body, each of which will conduct a certain amount of current with its resultant secondary field. Ore-bodies also occur at irregular angles or dips, and may be in thin sheets or veins. When it is recalled that *any* geophysical instrument operates on the combined *resultant* of all of these component forces, and that the nearest component has the greatest effect on the instrument, it can be readily seen how complex the correct interpretation of data may become from the one viewpoint of ore occurrence alone. To this may be added the topographic disturbances, variations in conductivity, tilting strata, and distortions to the primary and secondary fields, etc. Proper interpretation of geophysical data requires that due consideration be given the electrical, geological and tectonic, and mining data on each particular property.

Previous publications[5] have described the results obtained. Only a few types of survey data will be given in this paper to illustrate some interesting points regarding the application of the inductive process to geophysical exploration.

EXPLORATION MAY BE PROFITABLY PRECEDED BY GEOPHYSICAL EXAMINATION

Although it is well-recognized that geophysical methods alone are not a direct means of locating commercially valuable ore-bodies, such methods are of real economic value in outlining the development program of a property and securing information of value to the geologist regarding structural features and the location of the conductive mineralization. One valuable feature of the electrical methods usually is their ability to eliminate barren areas and to furnish information which the geologist and mining engineer may use to determine the most probable mineralized areas likely to contain commercially valuable

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[5] "The Radiore Method," E. H. Guilford, The Canadian Mining and and Metallurgical Bulletin, May 1928, pages 644 to 670.

"Fundamental Factors Underlying Electrical Methods of Geophysical Prospecting," J. J. Jakosky, *Engineering and Mining Journal*, Feb. 11 and 18, 1928.

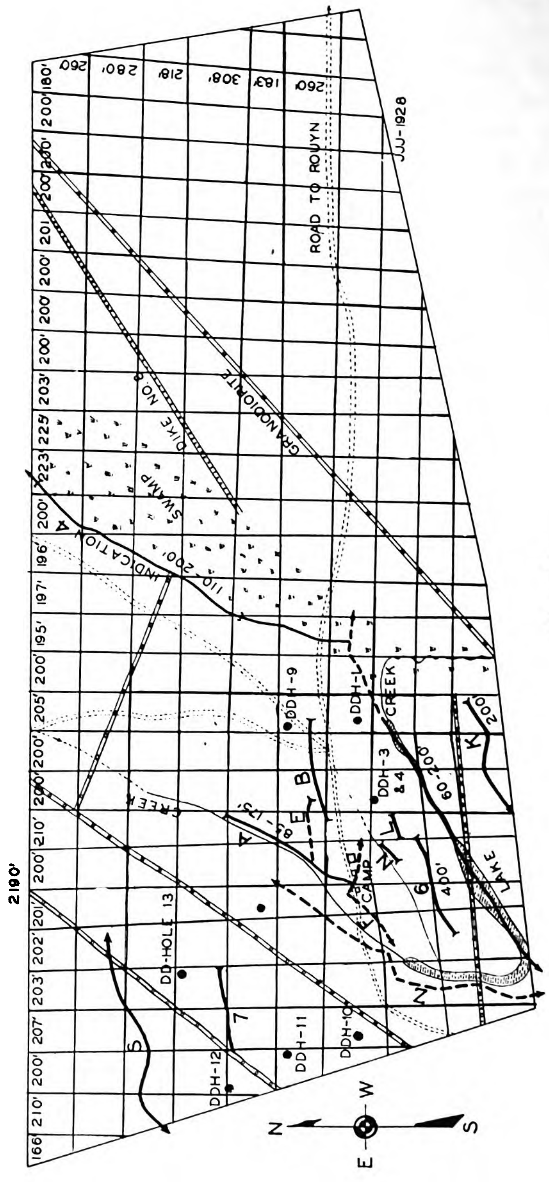
180'
200
200'
200
201
200'
200
200'
203'
225
223
200'
196
197'
195'
200
205'
200'
200'
210
200
201
202
203
207'
200'
210
166
2190'
260
280'
218'
308'
183'
260'
ROAD TO ROUYN
JJJ-1928
GRANODIORITE
DIKE NO. 8
SWAMP
INDICATION A
110-200'
CREEK
DDH-9
DDH-L
CREEK
DDH-3 & 4
60-200
K
200
B
E
A
85-135'
CAMP
F
L
N
6
400'
LAKE
N
DD-HOLE 13
7
S
DDH-11
DDH-10
DDH-12
N
W
E
S

Fig. 27

ore deposits. This may be illustrated by the exploration conducted by a Canadian Company in the Rouyn District.[6] This company had previously explored the region by a system of diamond-drilling, as shown in Fig. 27. The drill-cores gave some promise of mineralized zones to substantiate what little surface geology could be studied through the overburden typical of that district. This same property was later surveyed and the indications obtained, as shown on the map. It is safe to predict that had the electrical survey been conducted as one of the initial steps in the development of the property, the dia-

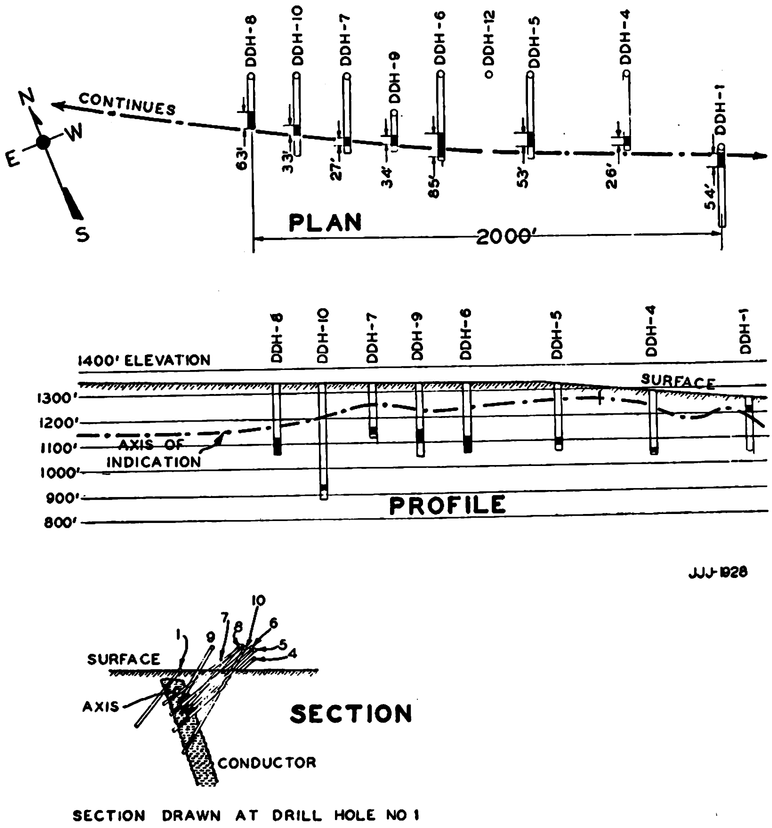

Fig. 28

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[6] Data from E. H. Guilford, of the Radiore Company of Canada, Limited.

mond-drilling could have been used in proving-up the indications and determining their value. Subsequent diamond-drilling on the indications has shown good mineralized zones.

CONTINUITY OF MINERALIZED ZONES

An interesting example[6] of the continuity of a mineralized zone may be seen in Fig. 28. This *indication* was located during a survey of a property in Des Meloizes Township, Quebec, Canada. The indication was found to be of proper electrical character to warrant further exploration work. This indication has now been checked over a distance of 2000 feet between drill hole No. 1 and No. 8, and every drill hole with one exception has cut a wide zone of sulphide mineralization. The conductor is a dipping sheet vein and possesses considerable depth.

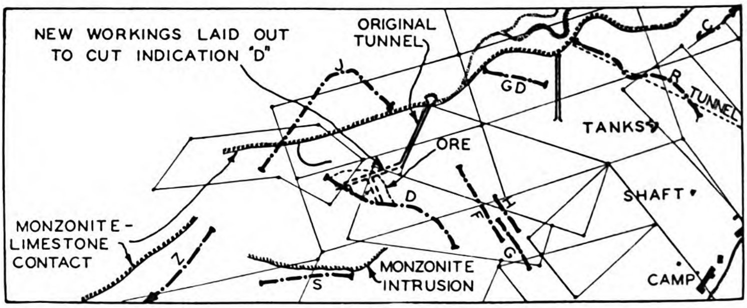
NEW WORKINGS LAID OUT TO CUT INDICATION "D"
ORIGINAL TUNNEL
J
GD
C
R
TUNNEL
TANKS
ORE
D
F
G
SHAFT
MONZONITE-LIMESTONE CONTACT
Z
S
MONZONITE INTRUSION
CAMP

Fig. 29

The "axis" of the conductor was determined by use of the "correction" curve. Note that this axis is close to the top of the conductor. As previously explained this is due to the portions of the conductor lying close to the surface having a much greater effect than the lower portions. As a general rule the effect of a differential conductive area will vary inversely as some power greater than the square of the distance to the direction-finding coil. An end view of the conductor and the drill holes are show. The mineralized portions of the drill hole are shown by solid black.

SPECIAL CASE WHERE SURFACE GEOLOGY COULD NOT GIVE INFORMATION

An illustration[7] of where surface geology failed to give any hint of a possible mineralized zone is illustrated by an indication

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[7] Data supplied by H. E. Olund, Geologist, The Radiore Co. of U. S.

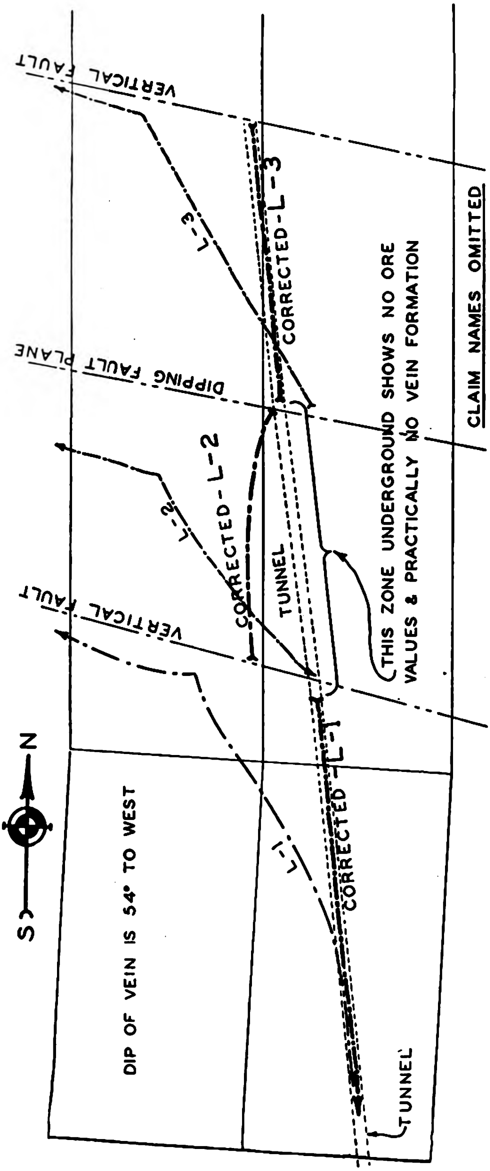
VERTICAL FAULT
L-3
CORRECTED-L-3
DIPPING FAULT PLANE
CORRECTED-L-2
L-2
TUNNEL
VERTICAL FAULT
THIS ZONE UNDERGROUND SHOWS NO ORE VALUES & PRACTICALLY NO VEIN FORMATION
CLAIM NAMES OMITTED
N
S
DIP OF VEIN IS 54° TO WEST
L-1
CORRECTED-L-1
TUNNEL

Fig. 30

located on an Inyo County, California, property, Fig. 29. This is a lead-silver district which has been mined intermittently since 1865. Experience from previous ore occurrences showed that the ore values always occurred in the limestone adjacent to the limestone-monzonite contact. A major conductor (named Indication D) was located during the electrical survey, but because of its distance from the contact as shown in the surface, considerable doubt existed as to its being of a special value. Because of its proper electrical characteristics, however, a cross-cut was driven, from the tunnel to Indication D, where it was found that a tongue of monzonite (probably an off-shoot from the main monzonite intrusion to the north) existed and the contact of this tongue with the surrounding lime had created conditions favorable for ore deposition. The high-grade oxidized ore associated with the sulphides was of excellent value, carrying approximately 21 ounces silver, and 33 per cent lead. The conducting sulphide zone consisted of galena, sphalerite, chalcopyrite and pyrite and is over one hundred and fifty feet in thickness.

CORRECTIONS FOR TOPOGRAPHY AND DIPPING CONDUCTOR

An example of the corrections which must be made for topography when working over dipping vein conductors is shown in Fig. 30 of a preliminary map taken from a Mexican Survey. The locations of the *indications* projected to the surface are shown by the heavy dotted lines. This property has been opened up by underground workings under the indications L1 and L3. No ore was encountered in the drift through one of these faulted blocks, and the electrical survey has shown the displacement of the vein in this block. Attention may be called to the apparent existence of a conductor on the continuation of the indication parallel with each fault. This probably is due to the faulted zone containing conductive material caused by the "drag" and wet clay gauge. The dip of the faulted plane, together with the dip of the vein of 54 degrees to the west, accounts for the displacement between the electrical indications and the surface fault projections. The conductive zones in this property are from 400 to 700 feet in depth. A diamond drill hole was driven from the tunnel to intercept the indication L2 and to establish identity with L1 and L3, but after penetrating about 30 feet encountered such a badly broken zone that further drilling was impossible. A crosscut will now be driven.

ACKNOWLEDGMENTS

The writer desires to acknowledge the material assistance and cooperation received from his associates of the Radiore Company, particularly to Messrs. A. B. Menefee, H. E. Olund, E. H. Guilford, M. Brenner, H. O. Walker, and R. N. Anderson. Especial acknowledgment is also made to Dr. D. A. Lyon and Mr. H. E. Head, of the Department of Metallurgical Research of the University of Utah, for assistance and help in the cooperative studies now being conducted on the electrical properties of various ores, etc.

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SOME STUDIES OF RADIO BROADCAST COVERAGE IN THE MIDDLE WEST*

By

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Summary—*Studies of the radio field intensities produced by broadcasting stations throughout the states of North and South Dakota and Minnesota have been made. From these studies and from information concerning the service rendered standards of service based upon field intensities have been suggested for daylight broadcasting in rural communities.*

Data concerning the distribution of radio sets throughout the three states in question have been gathered and ratios showing the percentage of families owning radio receiving sets have been calculated. The lack of correlation between this percentage and the availability of daylight broadcast service has been pointed out together with the fact that a large majority of radio listeners must rely entirely upon night-time reception from relatively distant stations. Some of the factors affecting reception from distant stations are mentioned. The apparent correlation between rapidity of fading and distance to the transmitting set is illustrated by fading curves.

QUANTITATIVE study of the electric field intensities produced by the radio broadcasting stations throughout a given territory yields valuable information, first, concerning the service areas of the stations studied and, second, the interference conditions which will be encountered in the use of receiving sets possessing given selectivity characteristics when subjected to known radio fields from two or more stations operating upon specific frequency assignments. This paper presents in part the results of certain studies made throughout Minnesota and adjacent states during the past two years with the view of determining some of the conditions affecting the service range of radio broadcasting stations located in the middle west.

The electric disturbances at a given receiving location which will cause sounds to be emitted from the loudspeaker of a radio receiving set will, for the purposes of discussion, be classified as follows, (1) a modulated signal such as would be produced by the operation of one broadcasting station, provided no other broadcasting station were in operation on a carrier frequency near enough to cause interference; (2) atmospheric disturbances; (3) interfering disturbances, produced by the operation of non-radio electrical devices and systems by man; (4) interfering disturbances produced by radio stations other than the one from which reception is desired.

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* Original Manuscript Received by the Institute, July 3, 1928.

The sensitivity of the great majority of radio receiving sets in use throughout the territory under consideration is such that it is not the actual value of the electric intensity produced by a radio broadcasting station which is of importance, but the ratio of that intensity to the intensity of interfering disturbances. If the ratio of the desired signal intensity to the intensity of the maximum interfering disturbance is sufficiently high and the sensitivity of the receiving set is sufficiently great, then satisfactory reception can be obtained by amplifying the signal to the desired volume. If this ratio is not sufficiently high, then satisfactory reception cannot be obtained. It is, of course, true that the permissible value which this ratio can possess will

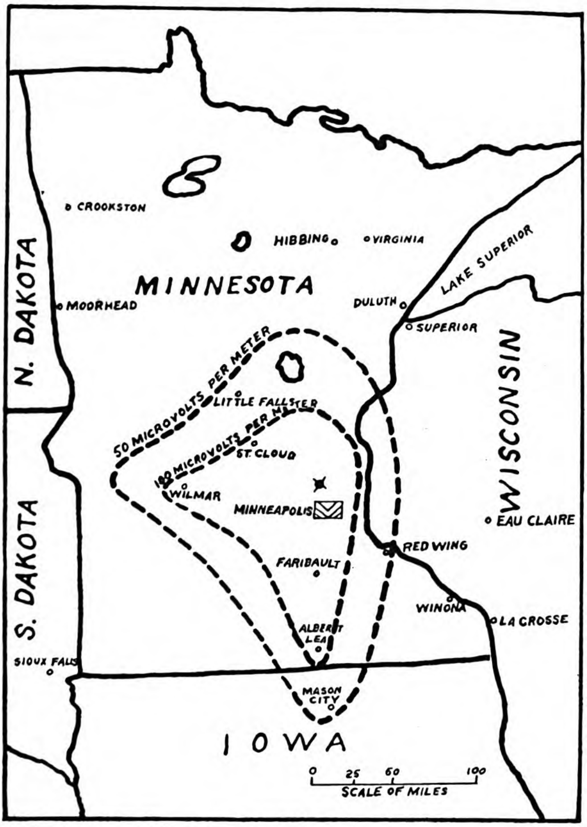

Fig. 1—Daylight Distribution of Radio Field from WCCO, 720 kc, 5000 watts.

depend to a certain extent upon the type of receiving set in use, type of signal collector, nature of interference, nature of program, etc., but these factors are of relatively small importance in comparison with some of the others involved.

The determination of the service area of a radio broadcasting station on the basis of the electric intensities produced throughout the territory under consideration involves, among other things, the establishment of a ratio of field intensity to static and interfering intensities which may be considered as determining the border line between satisfactory and unsatisfactory reception. It is also necessary to establish a standard of reliability which can be considered satisfactory. In the studies under discussion, these questions were not considered separately, but an attempt was made to set up daylight service standards by means of numerous observations utilizing a signal intensity measuring set throughout the area involved. Since fading and other peculiar transmission phenomena are of great importance in determining the night time range of stations, separate consideration must be given to night time conditions. These will be considered later.

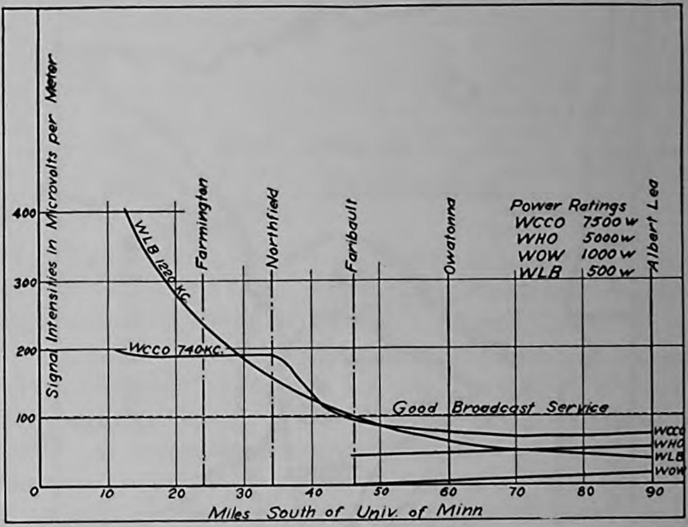

Fig. 2—Daylight Intensities Produced South of Twin Cities Along Minnesota Trunk Highway No. 1 by Radio Stations WLB (University of Minnesota), WCCO (Anoka), WHO (Des Moines), WOW (Omaha).

Both summer and winter daytime observations seem to substantiate the statement that, where rural or small town areas in the State of Minnesota are receiving daylight field intensities of the order of 100 microvolts per meter, the radio listeners inhabiting those areas feel that they are receiving *good broadcast service.* Where the field intensities produced are approximately 50 microvolts per meter, listeners feel that they are receiving *fair broadcast service.*

The author realizes that standards of service based upon field intensities as low as these are far from ideal and, in fact, far lower than those which must be established for thickly populated districts where interference from non-radio devices and other factors offer more serious limitations[1],[2]. However, unless we are willing to accept a standard of this order of magnitude for rural districts, we must somehow explain the almost every-day use of thousands of sets for daylight reception in localities which, on the basis of service standards higher than the one proposed, are receiving no broadcast service at all. Nevertheless, the owners of these sets are found to be reasonably well satisfied with the daylight service they are receiving when it is

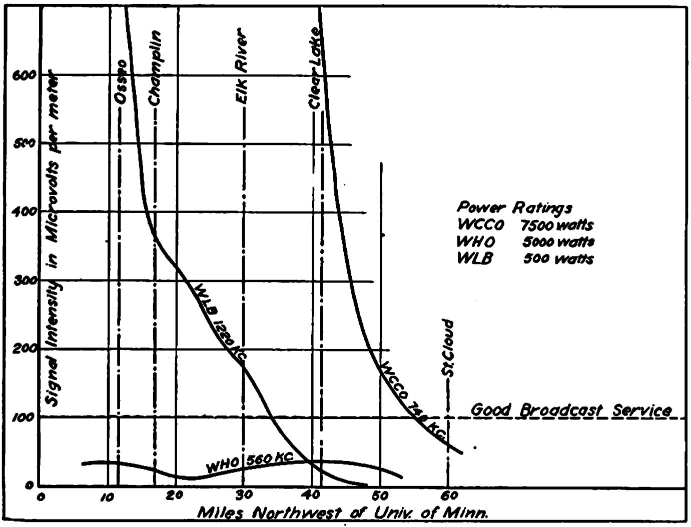

Fig. 3—Daylight Signal Intensities Produced Northwest of Twin Cities Along Minnesota Trunk Highway No. 3 by Radio Stations WLB (University of Minnesota), WCCO (Anoka), WHO (Des Moines)

supplied by signal intensities which equal or exceed that which has been proposed as a standard for "good broadcast service."

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[1] "Reduction of Interference in Broadcast Reception," by Alfred N. Goldsmith, Proc. I.R.E., 14, 575; October, 1926.

[2] "Radio Broadcast Coverage of City Areas" by Lloyd Espenshied, *Journal* A.I.E.E., 46, No. 1; January, 1927.

Fig. 1 shows the location of the 100- and the 50-microvolt-per-meter lines for WCCO, a 5000-watt station, as approximately determined by numerous observations made during the summer of 1926. Figs. 2 and 3 show daylight field intensities along lines running south and northwest, respectively, produced by WCCO and WLB (a 500-watt station operated by the University of Minnesota) during the fall and winter of 1927. Faribault and Albert Lea, which lie south of the Twin Cities, and St. Cloud, which is northwest, can be located on Fig. 1. In comparing field intensities produced south with those produced northwest it must be remembered that WCCO is located 20 miles north of the Twin Cities while WLB is located in the City of Minneapolis near the boundary line which separates that city from St. Paul.

Carefully prepared estimates place the number of radio receiving sets in Minnesota, North and South Dakota, February 1 1928 at 330,895. Of this number 116,425, or 38 per cent, are actually located on farms. 208,156, or approximately 62 per cent, are either on farms or in towns of less than 2500 population. (Residents of towns of less than 2500 population are considered as rural dwellers by the United States Census Bureau.)

Fig. 4 shows the distribution of receiving sets by counties in the three states under consideration. The upper figure in each county designates the number of sets on farms or in towns of less than 2500 population (sets owned by rural dwellers). The lower figure designates the number of sets owned by urban dwellers[3]. A study of Figs. 1 and 4 will show that of the 330,895 sets in the three states, only about 90,610 or 28 per cent lie within the "good" daylight broadcast service range of WCCO as defined earlier in this paper. Of this 90,610 only 23,810 or 27 per cent lie outside the Twin City Metropolitan district. For stations of lower power than 5000 watts, located in or near the Twin Cities, the percentage of sets served outside the metropolitan district will, of course, be still smaller.

Consideration of all of the broadcasting stations in or near the three states of Minnesota, North and South Dakota will show that, of the 330,895 sets, not more than 130,000 or 40 per

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[3] The data on the distribution of radio receiving sets in Minnesota, North or South Dakota contained in this paper were collected by J. O. Maland, of the *Dakota Farmer*, President of the Northwest Radio Trade Association.

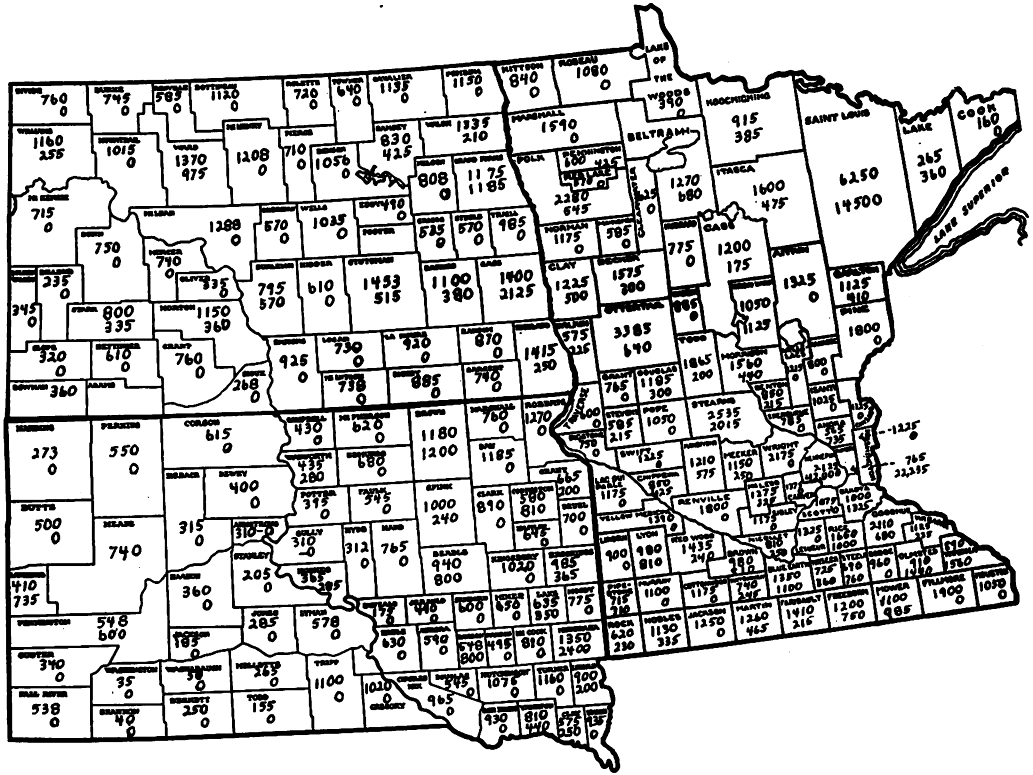

Fig. 4—Number of Radio Receiving Sets in Minnesota, North and South Dakota February 1, 1928. (Upper Figure—Sets Owned by Rural Listeners; Lower Figure—Sets Owned by Urban Listeners).

	Rural	Urban	Total
Minnesota	104,665	105,195	209,860
North Dakota	51,454	7,585	59,039
South Dakota	52,037	9,955	61,992
Total	208,156	122,735	330,891

cent lie within the "good" daylight broadcast service range of *any* station, even on the basis of as low a standard as 100 microvolts per meter. 200,000 or 60 per cent are on the basis of this low standard, getting no daylight broadcast service at all.

If the number of families and the number of receiving sets in a given community are known, then the ratio of the number of sets to the number of families, expressed in per cent, might be called the receiving set saturation factor for that community. Fig. 5 shows this saturation factor by counties for Minnesota, and North and South Dakota as obtained from Fig. 4 and from population figures for the various counties, assuming four people per family. As before, the upper figure refers to the rural population and the lower figure to the urban population.

It is evident from Fig. 5 that no apparent relationship exists between the saturation factors for the various counties and the availability of service on the basis of daylight field intensity standards. The natural question which arises is: "If the ratio of the number of receiving sets in a given community to the number of families does not depend upon the presence of adequate field intensities from broadcast stations, regardless of how low an intensity is considered as adequate, then upon what factors does it depend?" It is now quite universally accepted that radio broadcasting can fundamentally be of greater economic and social service to rural communities than to cities. From the high value of the receiving set saturation factor which exists in some of the counties, as shown by Fig. 5, which possess only a rural population, it seems, probably, that the value which this factor will possess in a given community is likely to depend more upon the need for broadcast service than upon its continuous availability.

The wide dissemination of receiving sets and the high value of the receiving set saturation factor throughout the three states, shown to exist in spite of absence of reliable daylight broadcast service, can only mean that night time reception of signals from stations at considerable distances is not only of value, but is all that the majority of listeners have available. Since large numbers of broadcast listeners in the central northwest, and similarly throughout large portions of the entire United States, have invested their money in receiving sets on the basis of distant night time reception and such reception only, schemes for the distribution of broadcasting stations which are intended to

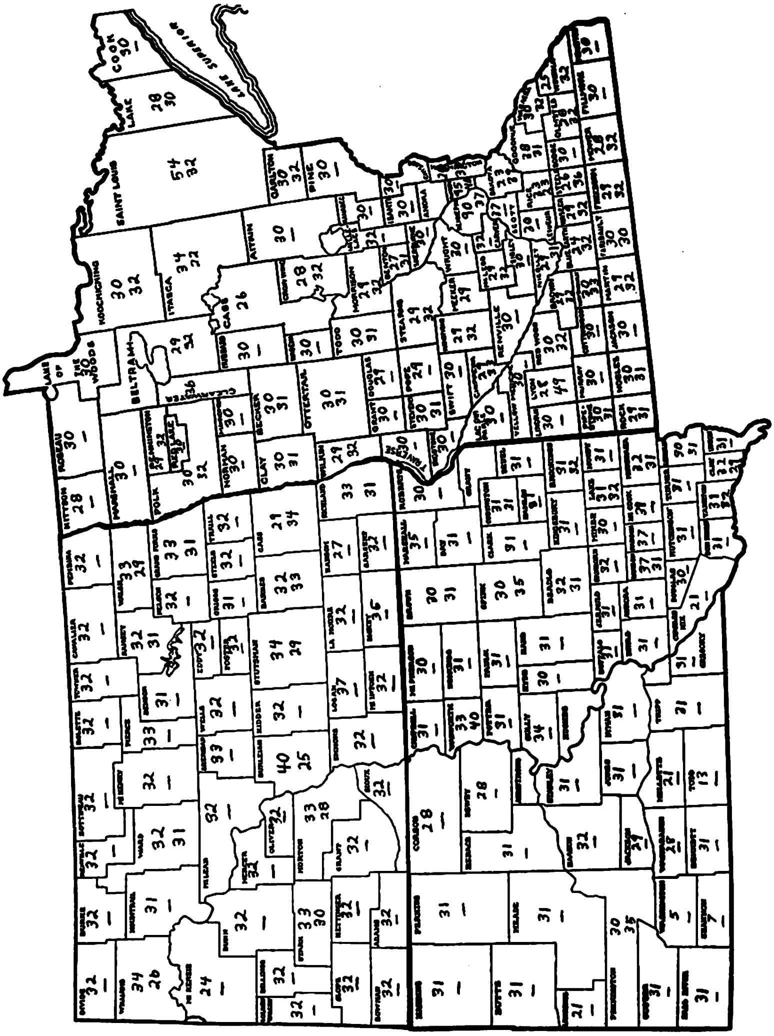

Fig. 5—Receiving Set Saturation Factors for Minnesota, North and South Dakota February 1, 1928 on the Basis of Four Individuals for Each Family. (Upper Figure Refers to Rural Population; Lower Figure Refers to Urban Listener).

deliver equitable broadcast service should give adequate consideration to receiving conditions which will exist several hundred miles from any broadcasting station.

An analysis of the conditions affecting night time reception in a region such as is under consideration is immeasurably more difficult than where high quality service range only is of interest. Absolute figures based upon field intensities alone do not tell the complete story. Atmospheric disturbances, non-radio electric disturbances, radio station interference and fading all become matters of great importance. On the other hand, the geographic separation between listener and broadcasting station becomes a factor of relatively minor importance. Often, reception from a particular station is more satisfactory at points 200 or more miles distant than at points lying between 50 and 200 miles[4]. While the scope of the investigations forming the basis for this paper was not such as to yield a great amount of information concerning these factors, important as they are and must continue to be to the broadcasting art for some time to come, they did, however, forcibly call attention to certain phenomena which are worthy of mention.

In general a station 1,000 miles away from a rural listener, free from heterodyne interference, will probably be of greater service to him than a station 200 to 500 miles away subject to such interference. Many instances are known where heterodyne interference from a 500-watt station 1,000 miles away has destroyed reception from a 5,000-watt station 20 miles away. It follows, therefore, that a broadcast structure designed to serve rural communities should, as far as possible, eliminate interference of this sort.

Since today and probably for a long time to come the rural listener must rely upon relatively low field intensities, it follows that any increase in power on the part of the stations serving him will improve his receiving conditions. If these stations are operating on clear channels, then such increase will not affect the general interference situation providing the stations involved are not located in thickly populated districts.

Of the remaining factors which limit distant reception, there is one, dependent upon the characteristics of the transmission medium, concerning which, as yet, far too little is known. Statistical data have repeatedly indicated that at night there is a zone which lies approximately between 50 and 200 miles of a broadcasting station in which reception is not as good as at points farther distant[4]. This condition has been found to exist in the territory under consideration. Sometimes the zone of poor reception extends closer than 50 miles and sometimes it extends further than 200 miles, depending upon weather conditions, frequency, and other factors. Under some conditions and with some transmitters, such programs as are received are so distorted as to be rendered useless. Increasing power at the transmitter does not seem to overcome the difficulty.

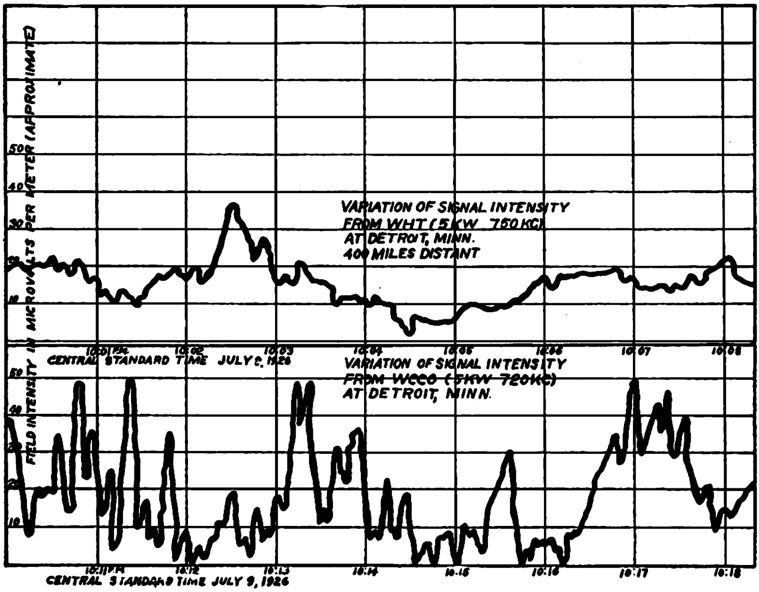

Fig. 6—Fading at 112 and 400 Miles.

Such evidence as has been collected seems to connect the condition described above with fading. As an illustration of a probable explanation, attention is called to Fig. 6 showing fading curves taken at Detroit, Minnesota in the summer of 1926. Data for the two curves were taken one after the other. The two stations involved were of the same make, type and power and were operated on frequencies separated by only 30 kilocycles. The field intensity from the station 112 miles away fluctuated much more rapidly than the field intensity from the station 400 miles away. Although the peaks of the curve for the nearby station rise somewhat higher than those for the more distant one and the average intensities are approximately the same, the program from the nearby station was not enjoyable while that from the distant station was. The listener was seriously disturbed by the rapid fluctuations in intensity produced by the former station while the slower variations of the latter were hardly noticeable. Had the observations been made at a point 80 miles from

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[4] "A Statistical Study of Conditions Affecting the Distance Range of Radio Telephone Broadcasting Stations" by C. M. Jansky, Jr., Technological Papers of the Bureau of Standards No. 297.

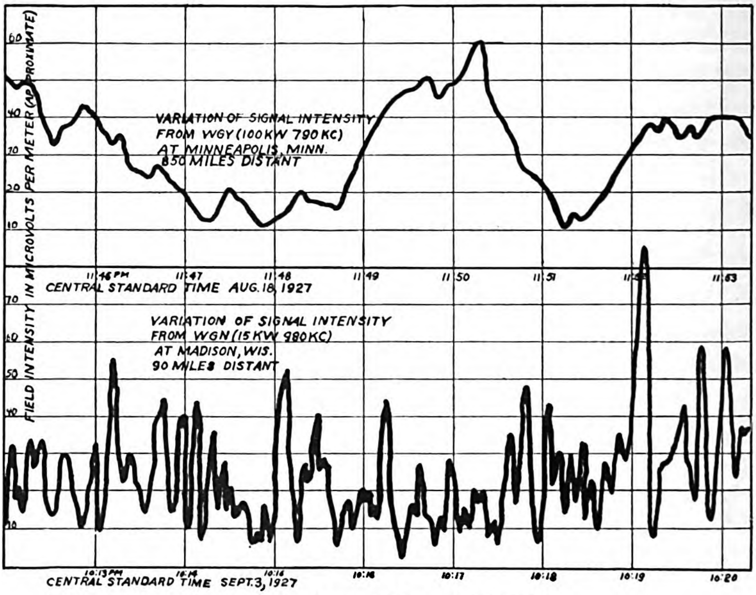

Fig. 7—Fading at 90 and 850 Miles.

the nearer station, the program from it would probably have been still less acceptable.

Fig. 7 shows curves obtained from two stations taken at different times and at different locations. Here the difference is still more marked. Numerous observations which the author has made upon stations operating throughout the broadcast spectrum lead to the conclusion that while the rapidity of fluctuation of intensity in the broadcast band at a particular time varies to some extent with the frequency in use at the broadcasting station, it varies to a much greater extent as the distance between the transmitter and receiving location is varied. The shorter the distance the shorter the average period. This, together with the effect of the rapidity of variations of intensity upon the sensitivity of the human ear to such variations, the author believes to be the main reason for the existence of the zone of poor reception.

When the more serious limitations imposed by heterodyne interference and lack of power have been removed, fading will probably be the most important remaining deterrent to the delivery of good night time broadcast service to listeners outside the high-quality service range of stations. The writer has believed for some time that the detrimental effects of fading as have been described might be largely overcome by radiating the modulated high-frequency energy from a plurality of locations which need not be so widely separated geographically as to make the scheme wholly uneconomic.

It seems, probably, that the most efficient utilization of a given broadcast channel could be obtained by operating a sufficiently large number of synchronized stations scattered throughout the entire nation, all carrying the same program. Such a system, if elaborate enough, would deliver adequate signal intensities everywhere and would also probably overcome the detrimental effects of fading. Aside from the technical difficulties to be encountered, the economic impossibility of inaugurating such a system in the immediate future should be obvious. The scheme proposed in the preceding paragraph is, however, somewhat different and certainly more workable from an economic standpoint at least.

The author wishes to express his indebtedness to Mr. R. R. Sweet, Chief Engineer of Station WCCO and to Mr. C. B. Feldman, Teaching Fellow at the University of Minnesota, who assisted in taking much of the data used in this paper, and to the management of WCCO whose interest in this subject made possible the studies of that station.

Volume 16, Number 10 *October, 1928*

A STUDY OF SHORT-TIME MULTIPLE SIGNALS*

BY
J. B. HOAG AND VICTOR J. ANDREW

Summary—*The echos or multiples of the direct signals from a transmitting station which arrive shortly after the main signal are due to direct-reflection from the Kennelly-Heaviside layer. Those arriving after considerable time (about one-seventh of a second) have travelled around the world. This paper deals with new, intermediate types of multiple signals which arrive at the receiving station from 0.01 to 0.04 second after the direct signal. Groups of these have been associated with paths to the regions of auroral activity, the magnetic and geographic north poles. A classification of these short-time multiple signals is given and resolved into the regular echos and a new wedge-shaped echo. The wedge or hangover multiples may be due to a plurality of paths resulting in successive echos of decreasing amplitude. Surprisingly large fluctuations of path difference of all short-time multiples occur in as short a time as one-quarter second.*

INTRODUCTION

THIS paper deals with radio signals which have traversed various paths between the transmitting and receiving aerials. The rays leave the transmitter in all directions both along the earth's surface and upwards. The first signals to arrive at the receiver have travelled in a straight line or along the shortest arc between the two stations. Later, "echos" or "multiples" of the direct signals come in (especially for very high frequencies) which have travelled upward and suffered reflection or refraction from regions of high electron concentration, the Kennelly-Heaviside layer, seventy-five to three hundred miles above the earth. Then there are multiple signals arriving at a considerable time later, about one-seventh of a second, which have made a complete excursion around a great circle of the earth. These latter may have travelled in the original direction of the direct signal or in the opposite sense and have been observed after three and four trips around the earth.

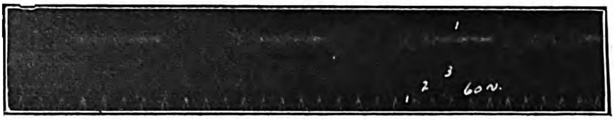

Fig. 1

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* Original Manuscript Received by the Institute, August 6, 1928.

Measurements were started on the first of this year with the intention of repeating Quack's[1] observations on these circumterrestrial signals and continuing their study. However, it was soon found that multiple signals often appeared after time intervals very much less than that expected for signals following great circle paths, but longer than for direct reflection from the Kennelly-Heaviside layer. In May, Taylor and Young[2] reported such short-time intervals in addition to circumterrestrial signals, and emphasized the need of further work at different geographical locations. This paper deals with such signals as observed in Ryerson Physical Laboratory at the University of Chicago during the first four months of 1928.

Fig. 2

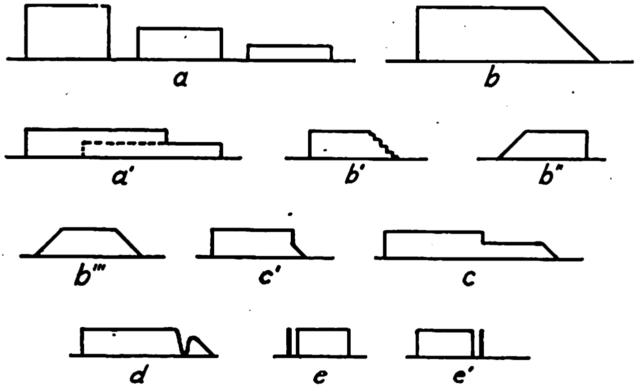

Fig. 3—Short Time Multiple Signals.

Records

Photographic records were made with a General Electric type EM oscillograph of signals from commercial stations operating on regular schedules between 10 and 13 megacycles. These frequencies are lower than those reported by Taylor and Young (14 to 20 megacycles) or Quack (around 18 megacycles). Interpretation of many of the records was difficult because of intense rapid fading of the types shown in Figs. 1 and 2. At times this fading was periodic and at other times it was extremely irregular. No attempt was made to study the phenomenon in detail.

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[1] Proc. I.R.E., 15, 341; April 1927.
[2] Proc. I.R.E., 16, 561; May, 1928.

After a large number of records were obtained it became possible to distinguish several different forms of distorted and multiple signals which appeared so consistently as to warrant classification. Idealized cases of these are shown in Fig. 3. Further study indicated that these could be reduced to two fundamental types, *a* and *b* of Fig. 3.

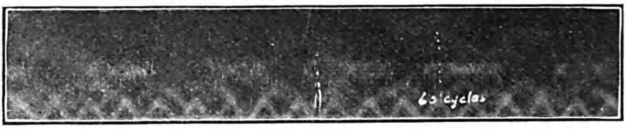

Fig. 4

The normal short-time multiple signals of type *a* are shown in Fig. 4, which is a series of dots from KEL, Bolinas, California, taken at 6:05 Central Standard Time, May 5th, on approximately 10.25 megacycles. It is significant that most of the dots reappear not only once, but twice, and those on the left three times. Furthermore, in the fraction of a second between successive dots a considerable variation in both time and intensity of multiple signals occurs, indicating great instability in the paths followed. The variation in time is seen from the detailed measurements of one record of KEL, given in Table I.

TABLE I

DOT NUMBER	INTERVAL IN SECONDS						
	Beginning of signal				End of signal		
1	—	—	—	—	0.0108	0.0195	0.0334
2	—	—	—	—	0.0093	0.0213	0.0320
3	—	0.0214	0.0435	—	—	0.0213	0.0433
4	—	0.0181	0.0435	—	0.0147	—	—
5	—	0.0302	0.0480	—	—	0.0193	—
6	—	—	0.0426	—	—	0.0170	0.0286
7	—	0.0276	0.0362	—	—	—	0.0331
8	—	0.0263	—	—	—	0.0232	0.0301

These may be roughly grouped into time intervals of 0.011, 0.022, 0.031, and 0.043 second. Records of the same station taken at various times give group values of 0.011, 0.012, 0.018, 0.028 0.029, 0.030 and 0.042 second. The average of these gives short-time multiple signals from this station of 0.011, 0.020, 0.029, and 0.042 second. These measurements include types *a* and *a′* since the latter is a special case of the former in which the main signal is of sufficient duration to cover the beginning of the multiple signal. Fig. 5 gives an example of this type. Similar records on AGB, Nauen, Germany, give averaged time intervals of 0.005, 0.006, 0.083, and 0.016 second.

The second fundamental type, *b*, which we shall call wedge signals, was at first thought to be a case of rapid fading. However, their continued appearance at times when fading was slight made it evident that a different phenomenon was involved. In a few cases a multiple signal consisted of one normal signal and one wedge-shaped signal. On these the entire length of the wedge-shaped signal, to the point of the wedge, appeared to be the length of the transmitted signal. The evidence was inconclusive, and it appeared more reasonable to believe the period of maximum signal strength equal to the length of the transmitted signal. If the latter is true, the signal shape may be due to a large number of multiple paths, giving a figure of type *b′*. This would correspond to the case mentioned by Heising[3] of transmission of a short jab of energy, and reception of a longer diffuse signal. Type *b* is shown in Fig. 6. The less frequent types *b″* and *b‴* are illustrated by Figs. 7 and 8 respectively.

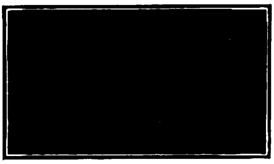

Fig. 5

Fig. 6

Type *c* is shown in Fig. 9. It is partially masked by a short extraneous noise of a wave form which often appears. Type *c′* occurs very frequently, and is illustrated by Fig. 10. *c* and *c′* are probably both combinations of types *a* and *b*. The principal signal is normal, and second signal wedge-shaped. Another variation of type *b* is type *d*, illustrated in Fig. 8.

Several signals of type *e* or *e′*, Fig. 11, appear in our records and in illustrations in Taylor and Young's article. Measurements of accompanying multiple signals indicate that the transmitted signal extended over the extreme length of the two portions of the signal.

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[3] Proc. I.R.E., 16, 75, January, 1928.

Measurements of the duration of the wedge or hangover of type *b* signals are given in Table II.

TABLE II

Station	Length in Seconds		
	Min.	Max.	Average
WIK	0.003	0.045	0.017
AGB	0.011	0.060	0.027
CJ	0.017	0.020	0.019
VIY	0.006	0.014	0.010
GBH	0.008	0.027	0.016
KEL	—	—	0.020

THEORY

Multiple signals must originate either in a plurality of paths of different length between the transmitting and the receiving stations, or in a plurality of signal velocities. Velocities differing as greatly as would be necessary to explain observed signals are too improbable to warrant consideration. Attempts to explain the origin of multiple signals must be based on the existence of several paths over which the signal arrives, travelling at approximately the velocity of light. To account for the measurements discussed in this paper the lengths of the paths must vary over wide limits in a fraction of a second. It is, of course, not inconceivable that different phenomena are involved in indirect paths of different lengths. Since there is no other evidence of reflecting or refracting surfaces at great distances above or below the earth's surface attention must be directed to the possible paths approximately parallel to the earth's surface.

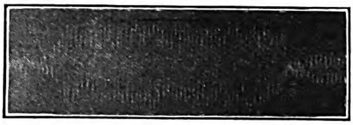

Fig. 7

Taylor and Young have suggested two possibilities, reflection from regions of high electron concentration near the earth's magnetic pole and scattered reflection from rough country beyond the receiving station. The first of these checks well with the 0.020 time intervals from KEL, but we are still confronted with the 0.011, 0.029, and 0.042 values. The Aurora Borealis is known to be most prominent above North America in latitude 60 deg. north and in the north Atlantic, falling off to the north and to the south. Its height ranges from 50 to 100 miles above the earth's surface for the lower strata of light and 75 to 200 miles for the upper. Further, brilliant Auroras accompanying magnetic storms and the Auroral rays follow the lines of the earth's magnetic field. Rough calculations on the time intervals, 0.011 from KEL and 0.005-0.006 from AGB for indirect paths to 60 deg. north check as well as could be expected. It is highly desirable that experiments on the direction of the incoming direct and multiple signals be performed and an attempt be made to obtain a correlation with the Aurora or magnetic storms.

Fig. 8

Fig. 9

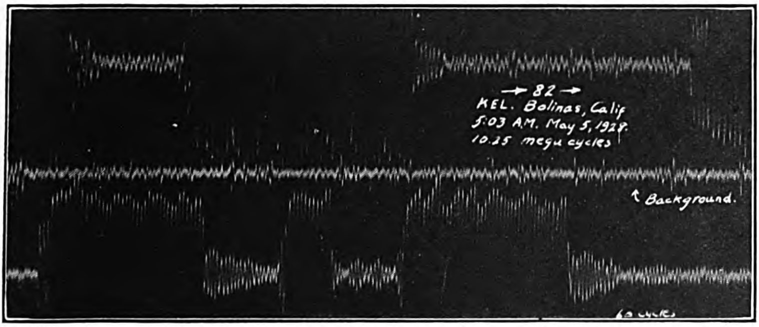

Fig. 10

The 0.042 values indicate a path as far north as the geographic pole. It is suggested that, since the arctic regions are in comparative darkness for many months of the year, the Kennelly-Heaviside layer is considerably elevated over these regions, like an inverted bowl, and that the rays are reflected around its inner surface. These signals would be expected to be rarer during the arctic summer months.

The Kennelly-Heaviside layer, normally parallel to the earth's surface, has a vertical component along the twilight zone. For signals travelling at a certain elevation and encountering this at a certain angle, it would serve as a source of horizontal reflection or refraction. Such deflection, if large, might produce indirect paths. Friis[4] has reported a large and rapidly changing error in high-frequency direction finding when the twilight zone lies between the transmitting and the receiving stations.

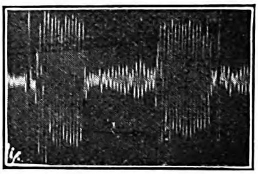

Fig. 11

In considering the wedge type signals, and especially the type *b′*, it is proposed that these consist of the main signal followed by a large number of multiple signals close together and of decreasing amplitude. This would require a band of indirect paths or echelon reflection. Since the time duration of these hangover signals is different, even in one-quarter of a second, the step-like reflector must be moving rapidly. Measurements on KEL check roughly for reflections from 60 deg. to 80 deg. north latitude. Here we are again reminded of the intense Auroral or magnetic action. If such a correlation be established these multiple signals will serve to study the nature of Aurora as well as the propagation of radio signals.

It is hoped that additional information may be obtained, especially from stations farther north, including the direction of reception and angle of polarization of the received signals.

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[4] Proc. I.R.E., **16**, 658; May, 1928.

Volume 16, Number 10 *October, 1928*

A NEW METHOD FOR DETERMINING THE EFFICIENCY OF VACUUM-TUBE CIRCUITS*

BY

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Summary—*A new method for determining efficiency of vacuum-tube circuits at both high and low frequencies has been described. This method consists of the application of a surface pyrometer which indicates the temperature of the glass walls of a vacuum tube. Readings are made of the temperature of the tube walls when the circuit is oscillating together with notation of d.c. plate watts input and then oscillation of the tube system is stopped and the plate input watts changed until identical pyrometer reading is obtained. The difference in wattage between the two measurements represents the radio-frequency output power.*

The simplicity of the method and the short time required for the measurement recommends it for serious consideration where air-cooled tubes are employed.

VARIOUS methods for determining the efficiency of vacuum-tube amplifying and oscillating circuits have been used to date with some success, especially in the low-frequency band. The advent of high-frequency communication systems in the band between 3000 and 20,000 kc and studies of vacuum-tube circuit efficiencies in this band have shown that certain methods for determining tube circuit efficiencies are not reliable. It is the object of this paper to describe a new method which has been found to be reliable for any frequency range when applied to circuits which employ the air-cooled vacuum tube.

Our experience with the commonly-known resistance variation and the optical pyrometer methods in the high-frequency band has proven them to be unsatisfactory. In the resistance variation method, where we determine the tank circuit resistance and, knowing the current flow, estimate the watts dissipated, we are very dubious of our resistance standards. These standards are most likely accurate at low frequencies but are not so in the high-frequency band. This condition can be attributed to the presence of appreciable inductance and distributed capacity in these standards which, if not compensated for, will introduce large errors. The general use of low resistance tank cir-

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* Original Manuscript Received by the Institute, August 15, 1928.

cuits with the appreciable resistance of tuning condenser and thermoammeters do not permit checking of measurements to less than ten per cent. The precautions to be observed when using this method at high frequencies and also the time required for such a measurement do not recommend it for quick and accurate work.

The optical pyrometer method is another means for obtaining tube circuit efficiency, but is not preferred for precision work with air-cooled tubes. In this method the optical pyrometer is employed to observe the temperature of the plate of the vacuum tube. A run is made with the tube in the non-oscillating state and the temperature of the plate is noted for different plate watt dissipations. The circuit is then placed in the oscillating state and the temperature of the plate is again obtained by use of the optical pyrometer. The plate watts input is noted for the oscillating condition, and from this is subtracted the watts dissipation represented by the pyrometer reading. The remaining number represents the watts which have been converted into radio-frequency power.

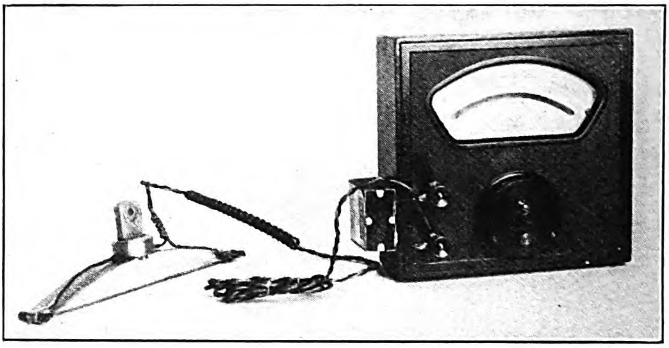

Fig. 1

This method is easier to apply than the resistance variation method but has its limitations. These limitations are variation of temperature over the surface of the plate for the direct-current dissipation measurement and a change in color for the same dissipation when the tube circuit is in the oscillating state. It is a common occurrence to observe a definite part of the plate that is of a cherry red color while other areas are of a dull red or black color when the tube is dissipating a certain wattage during the direct-current test, but as soon as the tube circuit starts oscillating and the plate is dissipating the same number of watts the color pattern spreads out and no longer shows the definite color portion. This spreading out is more pronounced as the frequency generated by the circuit is increased, particularly when operating in the high-frequency range. This change in color for the two conditions introduces a considerable error at high frequencies and therefore eliminates the system from consideration for precision measurements.

Another serious difficulty with this method, especially when the tube circuit efficiency is high, is the lack of sufficient color on the plate to permit an accurate pyrometer reading. From the foregoing statements it can readily be seen that only at excess power ratings or inefficient tube circuit conditions is it possible to use this method.

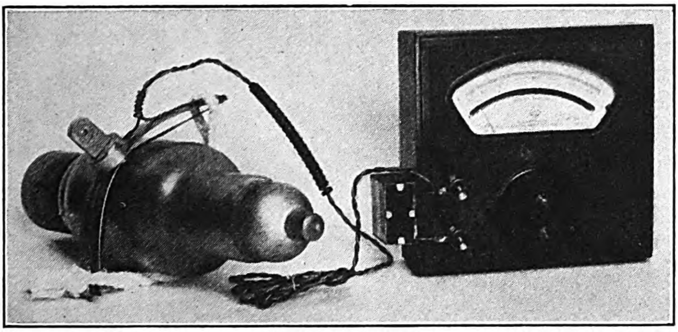

Fig. 2

Having determined the limitations and difficulties encountered in these two methods it became imperative that a more simple and reliable method be developed for measurement of the efficiency of air-cooled tube circuits. An opportunity for the solution of this problem was presented in the use of a Cambridge surface pyrometer. This pyrometer was designed primarily to measure the temperature of rolls and other cylindrical or flat surfaces which are employed in the manufacture of paper, rubber, and textile materials.

The type employed in our work is shown in Fig. 1. In this figure a flat strip thermocouple is illustrated with flexible leads connecting it to a millivoltmeter. The thermocouple strip is stretched between the ends of a heavier metal strip and insulated from it by hard rubber end pieces. A paper condenser of 1 μf capacitance is shunted across the millivoltmeter for the purpose of bypassing any radio-frequency currents that may be picked up by the thermocouple and leads.

Figs. 2, 3, and 4 show the pyrometer and the method of connecting it to the 250-watt, 50-watt, and 75-watt (shield-grid) tubes. This type of pyrometer measures the temperature of the glass walls of the tube and in part indicates an average or integrated value of temperature, while the optical pyrometer measures the temperature of the particular part of the plate upon which it is focused. The heat delivered to the thermocouple from the plate of the tube has to pass through the evacuated space between the plate and the tube wall and then through the glass wall and enroute it is averaged out and represents, indirectly, the heating effect of the entire plate.

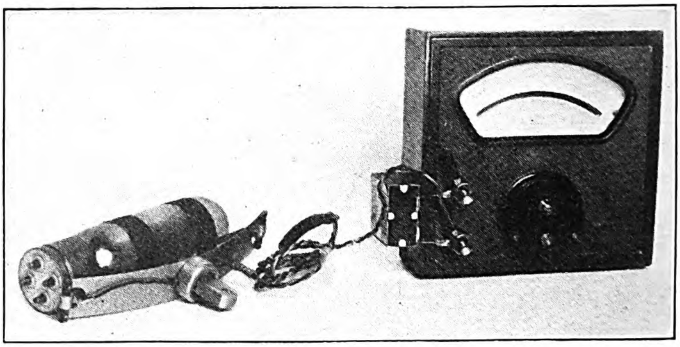

Fig. 3

For precision measurements it is advisable to explore the entire surface of the tube and find a place where there is minimum temperature difference between readings taken with the tube circuit in the oscillating and non-oscillating state. There will be a difference in temperature for these two conditions on certain parts of the tube surface, but there are a number of places on the tube surface where there is no perceptible temperature difference and at one of these places the pyrometer is secured.

The same method of determining the output of the tube circuit is employed with the surface pyrometer as was employed with the optical pyrometer; namely, to obtain a run showing temperature difference for different direct-current plate dissipations and then with the circuit in the oscillating condition note the input watts to the plate and subtract from this the dissipation wattage represented by the pyrometer reading with the remaining number of watts equal to the radio-frequency output.

The simplicity of the method and the ability to repeat the experiments and obtain results that check to less than 10 per cent recommends this new method for consideration.

There are a number of applications of this method for important measurements, among which are the following:

Antenna resistance
Tank circuit resistance
Condenser resistance

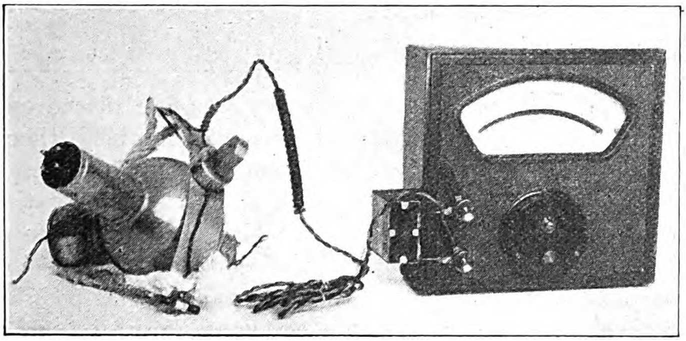

Fig. 4

Choke coil losses
Inductance coil resistance
Output power in antenna or tank circuit
Efficiency of amplifier tube circuits when singling, doubling, or tripling frequency
Efficiency of oscillator systems

An example of the various uses to which this system can be applied may be noted from the measurements which are described in the following paragraphs.

CRYSTAL-CONTROLLED OSCILLATORS

An investigation was made to determine the efficiency of the Navy standard crystal-oscillating circuit[1] and also the push-pull, crystal-oscillating circuit. The Navy standard circuit employs a single 50-watt tube with reduced plate voltage and a parallel feed plate circuit. The push-pull circuit employs two 50-watt tubes operating at same plate voltage as the single tube circuit with the push-pull principle, wherein the crystal is connected between the two grids and the plates are connected to the opposite ends of the tank circuit.

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[1] A. Crossley, Proc. I.R.E., 15, 9; January, 1927.

There was some doubt as to the efficiency of the push-pull crystal-oscillating circuit, particularly the question of whether or not the output was equal to twice that obtainable with the single-tube circuit. Table I shows definitely that the output obtained with the push-pull circuit is double that of the single-tube circuit.

TABLE I

CIRCUIT	INPUT (Watts)	PYROMETER READING	DISSIPATION (Watts)	OUTPUT (Watts)	EFFICIENCY
Single tube oscillator	62.	447 deg. F	41.	21	34 per cent
Push-pull tube No. 1	73.5	452	50.5	23	31 per cent
Push-pull tube No. 2	73.	448	52.	21	29 per cent

In this measurement the pyrometer was first placed on one tube of the push-pull circuit and data was obtained, then the pyrometer was shifted to the second tube and the experiment was repeated. In this way it was possible to check the operation of each tube, thus obtaining the data which is shown in Table I. Repeated checks were made on the efficiency of the two circuits in the band between 2000 and 4000 kc, and it was possible to repeat such measurements with a maximum error of 5 per cent.

VACUUM-TUBE AMPLIFIERS

Measurements were made on the efficiency of the CG-1860 250-watt and the Westinghouse AT-671 250-watt (shield-grid) tube when operated as an amplifier in the 4000- to 4500-kc band. Figs. 5 and 6 show graphically the efficiency and output obtainable with these two types of tubes.

In these measurements it was desired to determine whether the shield-grid tube was as efficient as the three-electrode tube and in addition what was the relation between excitation and negative *C* grid voltages for maximum output and efficiency. These tubes obtained their excitation voltage from a push-pull, crystal-controlled master oscillator. The three-electrode tube amplifier stage was balanced by resort to the Hazeltine method of neutralization in order that a fair comparison could be made with the shield-grid amplifier.

In computing the efficiency of the shield-grid amplifier, the dissipation in the shielding grid was considered. The variation of negative *C* voltage produced a variation in shielding grid dissipation from 2 to 32 watts which values were considered when computing efficiencies.

Referring to Figs. 5 and 6, the solid lines represent tube-circuit output while the dotted lines indicate efficiency. A study of these figures show, first, that the three-electrode tube is a more efficient amplifier than the shield-grid tube and secondly greater efficiency and output is obtainable with greater excitation voltage. The ratio of excitation peak voltage to negative *C* voltage for maximum efficiency is approximately two to one.

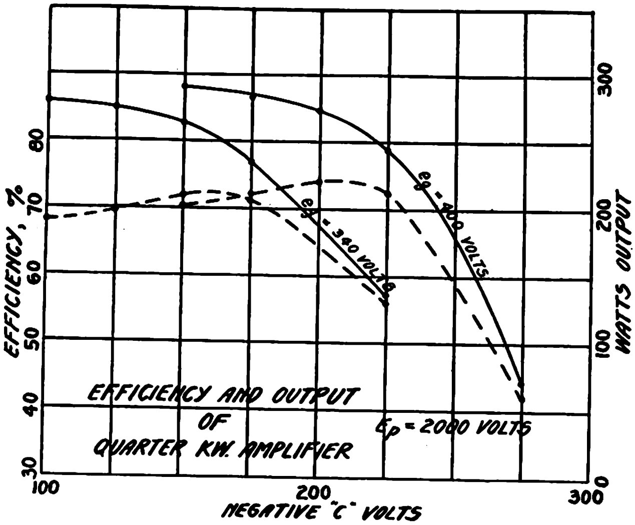

Fig. 5

In this experiment a vacuum-tube voltmeter was employed to determine the excitation voltage to the grid of the amplifier tube. Due to this method of measurement only peak voltages can be obtained and consequently all values of excitation voltage quoted in this paper are peak values.

ANTENNA RESISTANCE

The measurement of the resistance of antennas operating in the band between 4000 and 20,000 kc presents a problem which cannot be successfully solved by the resistance variation method. This is due to the fact that such antennas cannot be duplicated very readily by dummy circuits which employ lumped inductance and capacity. These high-frequency antennas are not loaded as are the low-frequency antennas, but have a considerable amount of radiation resistance which is many times the resistance represented by the d.c. and dielectric losses.

Generally, these antennas are of such length that they are a whole wavelength or a fraction of a wavelength long. Best practice at present time calls for antennas which are one-quarter, one-half, or three-quarter wavelength long according to type of radiating system employed.

The pyrometer method is well-adapted to the measurement of power which is delivered to the antenna system, and if proper precautions are made to note the current at the current node of the antenna, it is possible to measure the antenna resistance.

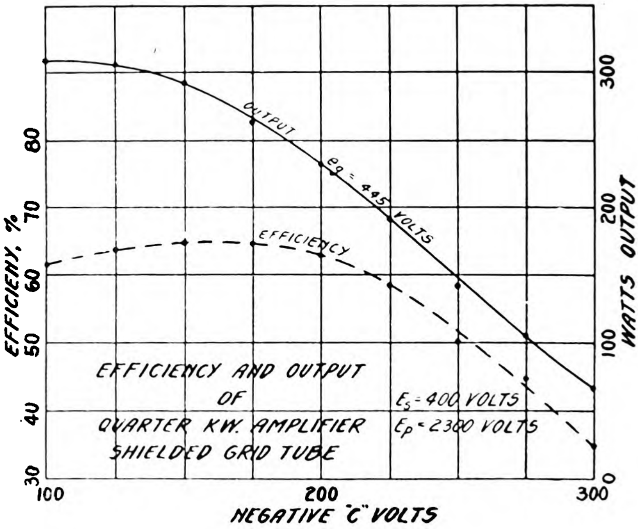

Fig. 6

To obtain this data with the pyrometer, we first compute the output of the tube system in the same manner which was described in the preceding chapters having the antenna connected to the tube system. In this measurement the current in the tank and antenna are recorded, after which the antenna is disconnected.

The next step is to retune the tank circuit to resonance and reduce the power until the current value obtained in the tank circuit is the same as when the antenna was connected. Now measure the output of the tube and subtract this value from that which was obtained with the antenna connected. The remaining power represents the antenna output. The antenna resistance is computed by dividing the antenna power by the square of the antenna current.

TABLE II

TOTAL OUTPUT (watts)	TANK CIRCUIT (watts)	ANTENNA (watts)	ANTENNA RESISTANCE
283	47	236	49.7 ohms
159	32	127	49.7 "
302	66	236	49.7 "
362	47	315	56.1 "
311	41	270	52.4 "
254	66	188	45.5 "
333	46	287	55.8 "
	Average Antenna Resistance		51.3 ohms

Table II indicates the accuracy of the new method and cites measurements made over a period of ten days on a quarter-wave vertical antenna which was operating on 8000 kc.

Volume 16, Number 10 *October, 1928*

SOME PRINCIPLES OF GRID-LEAK GRID-CONDENSER DETECTION*

By

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(Stanford University, California)

Summary—*The action taking place in the grid-leak, grid-condenser method of detection is reduced to an equivalent circuit consisting of the grid-leak condenser impedance in series with the dynamic grid resistance. The effect of applying a signal voltage to the detector, as far as the rectified grid current is concerned, can be represented by the introduction in the equivalent circuit of a fictitious voltage that is determined by the signal, and by a single tube constant called the voltage constant. The voltage drop produced by this fictitious voltage across the grid-leak condenser impedance is the change of grid potential resulting from the detector action. A bridge method of measuring all detector constants is described. The effects of grid leak and condenser sizes are discussed, and the factors involved in the selection of conditions for best detection with telephone and telegraph signals are considered.*

Introduction

AT THE present time there is no way of treating the phenomenon of detection of radio signals by the grid-leak grid-condenser method that is satisfactory for engineering purposes. The theory of detectors has been developed in a more or less complete form by various investigators,[1,2] but there has been no simple way of visualizing the quantitative relations existing in the detection process, and of measuring the detector coefficients. As a result the computation of detector performance is practically never undertaken.

PROCESS OF GRID-LEAK GRID-CONDENSER DETECTION

In grid-leak grid-condenser detection, the circuit for which is shown in Fig. 1, the applied signal voltage is rectified by the non-linearity of the grid-filament circuit, producing a rectified

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* Original Manuscript Received by the Institute, July 31, 1928.

[1] "A Theoretical and Experimental Investigation of Detection of Small Signals," by E. L. Chaffee and G. H. Browning, PROC. I.R.E., **15**, 113; February, 1927.

"On Optimum Heterodyne Reception," by E. V. Appleton and Mary Taylor, PROC. I.R.E., **12**, 277; June, 1924.

"The Rectification of Small Radio-Frequency Potential Differences by Means of Triode Valves," by F. M. Colebrook, *Experimental Wireless*, **2**, 865; 1925.

[2] Since this manuscript was first drafted there has appeared an article by Stuart Ballantine, "Detection by Grid Rectification with the High-Vacuum Triode," PROC. I.R.E., **16**, 593; May, 1928.

grid current that must necessarily flow through the grid-leak grid-condenser impedance and produce a voltage drop. This voltage drop then affects the plate current of the tube by ordinary amplifier action.

The amplitude of the rectified grid current depends upon the amplitude of the signal voltage. When the signal voltage is a modulated alternating-current voltage, the rectified grid current varies in amplitude at the frequency of modulation. Thus, when the radio-frequency signal voltage is modulated at a frequency of 1000 cycles, the rectified grid current pulsates in amplitude at 1000 cycles, and accordingly contains a direct-current part with superimposed alternating-current components having a fundamental frequency of 1000 cycles.

The problem of grid-leak condenser detection is centered around the determination of the change of detector grid potential produced in the detecting or rectifying process. The effect which a known change of grid potential produces in the plate circuit is an amplification problem and can be determined from the theory and equations of the vacuum-tube amplifier. The sensitiveness of the detector, that is, the change of grid potential produced by a given signal voltage, depends upon the curvature of the grid-voltage grid-current characteristic at the operating grid potential, and upon the impedance offered to the rectified grid current by the grid-leak condenser combination.

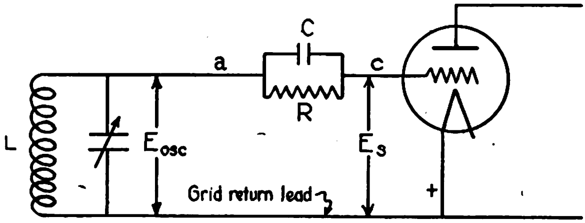

Fig. 1—Circuit for Grid-Leak Grid-Condenser Detector.

In radio-telephone and telegraph reception it is the audio-frequency components of the rectified grid current that convey the intelligence, and the grid-condenser grid-leak combination should accordingly offer a high impedance to audio-frequency currents. The impedance of the grid-leak condenser combination depends upon frequency, however, and becomes less as the frequency is increased. The detector is accordingly less sensitive for high modulation frequencies. The result is distortion in which the high notes are discriminated against.

Equivalent Circuit of the Grid-leak, Grid-condenser Detector

The change of grid potential which a given signal will produce can be analyzed with the aid of the equivalent circuit of the grid-leak condenser detector shown in Fig. 2. This equivalent circuit is based on the following law of detectors:

The rectified grid current produced by application of a small signal voltage (such as 0.05 volts effective or less) to the grid of a detector tube is exactly the same current that would be produced by a series of suitable generators acting between the grid and filamen in series with the grid-filament resistance of the tube. There is one such generator for each frequency component of the rectified current. The proper amplitude and phase of each generator depends only upon the operating grid potential, and the nature of the signal, and is given in Table I. The current produced by these fictitious generators if actually present in the grid circuit is the rectified current that the applied signal voltage produces. A proof of this proposition is given by Carson.[3]

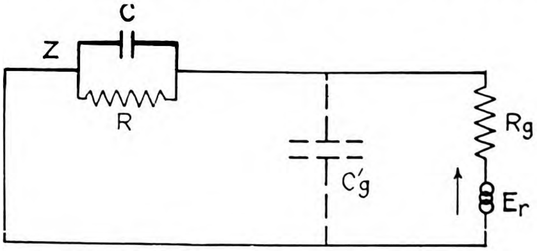

Fig. 2—Equivalent Circuit of Grid-Leak Grid-Condenser Detector.

The generators that can be assumed acting between the filament and grid in series with the dynamic grid resistance to produce the rectified grid current have no actual existence. It is merely that the effect of applying a signal voltage to the grid is the same as though these fictitious generators actually were present, and as though they and not the signal voltage were the

[3] The writer discovered this proposition in the latter part of 1926, and then found that it had been derived by three previous investigators, namely, Chaffee and Colebrook in the references cited, and John R. Carson, "The Equivalent Circuit of the Vacuum-Tube Modulator," Proc. I.R.E., 9, 243; June, 1921. Carson seems to be the first, although his application of the idea was to plate-circuit modulation.

forces really producing the rectified grid current. The voltage developed by this series of fictitious generators that represents the rectifying effect of the grid upon a signal voltage can conveniently be called the *rectified grid voltage,* and will be represented by the symbol E_r.

The rectified grid voltage E_r has a number of components, one for each component of the rectified grid current. Table I gives the frequency, amplitude, and phase of the different components that are present with various types of signal voltages. The rectified grid voltage E_r always contains a direct-current component, and when the signal voltage consists of several voltages of different frequencies (as a modulated wave which has a carrier wave and two side bands) the rectified voltage has in addition to the direct-current component a component of every possible difference frequency that can exist with the signal frequencies present.

The results of Table I are special cases of a general formula. To obtain the rectified grid voltage in the general case one writes down the signal voltage e_s as a function of time, squares this function, and divides by v. The low-frequency terms of the result are the components of the rectified grid voltage.

The rectified grid potential E_r is determined only by the signal voltage, and a tube constant v that will be discussed in the next section.

The size of grid condenser and grid leak has no effect on the amplitude of the rectified voltage E_r except insofar as the grid-leak resistance affects the operating grid potential. The current that the rectified voltage produces in the grid circuit, the rectified grid current, therefore depends upon the impedance of the grid-leak condenser combination, and consequently upon the size of grid condenser and grid leak. Each component of the rectified grid voltage acts by itself in producing its component of rectified grid current, independently of the presence or absence of other components of rectified grid voltage and current.

TABLE I

FORMULAS FOR RECTIFIED GRID VOLTAGE

1. Unmodulated radio signal voltage of crest value E_s, having equation $e_s = E_s \sin \omega t$.

   Components of rectified grid voltage:

   Direct-current component $= +E_s^2/2v$

2. Two superimposed alternating-current voltages of amplitudes E_1 and E_2, and frequencies f_1 and f_2, respectively, having equation $e_s = E_1 \sin 2\pi f_1 t + E_2 \sin(2\pi f_2 t + \phi)$
   Components of rectified grid voltage:
      Direct-current component $= +(E_1^2 + E_2^2)/2v$
      Component of frequency $f_1 - f_2 =$
         $+(E_1 E_2/v)\cos[2\pi(f_1 - f_2)t - \phi]$
3. Modulated wave with carrier crest amplitude E_m and degree of modulation k, having equation $e_s = E_m(1 + k \sin qt)\sin \omega t$
   Components of rectified grid voltage:
      Direct-current component $= +(E_m^2 + k^2 E_m^2/2)/2v$
      Modulation frequency component $= +(E_m^2 k/v) \sin qt$
      Double modulation frequency component $=$
         $-(k^2 E_m^2/4v) \cos 2qt$

Notation and formulas:

$R_g = dE_g/dI_g =$ dynamic grid resistance
$v = 2R_g/(dR_g/dE_g) =$ detector voltage constant
$E_r' =$ a component of rectified grid voltage.
$Z' =$ vector impedance of grid-leak condenser combination to component of rectified grid voltage E_r'

$$\left.\begin{array}{l}\text{The change of grid potential}\\ \text{produced by component } E_r'\end{array}\right\} E = E_r' \frac{Z'}{R_g + Z'}$$

Note that all amplitudes given in this table are crest values.

The capacity C_g' shown in dotted lines in the equivalent detector circuit of Fig. 2 is the effective grid-filament input capacity of the tube at the frequency of the rectified voltage component being investigated. This capacity depends upon the plate-circuit impedance of the tube, and in general its value changes with frequency. As C_g' is in parallel with the grid condenser C' it has the effect of making the grid condenser correspondingly larger. Strictly speaking, the effective grid-filament capacity C_g' should be shown shunted by the grid-filament input resistance, but for practical purposes the simpler arrangement illustrated in the figure is perfectly satisfactory.

The change of grid potential that is produced by the application of the signal voltage to the grid can be readily determined with the aid of the equivalent detector circuit of Fig. 2 and the formulas of Table I. This change of grid potential is simply the voltage developed across the grid-leak condenser impedance by the rectified grid voltage E_r acting in the equivalent circuit.

While the rectified voltage usually has a number of components, ordinarily only one of them is of practical importance. Thus in detecting a modulated signal voltage, the chief interest is in the component of the grid-potential variation that takes place at the modulation frequency. The change of grid potential for any such frequency component can be obtained as follows: Let E_r' be the component of rectified voltage of the desired frequency, as obtained from Table I, and let Z' be the impedance offered to this frequency by the grid-leak condenser combination. Then calling ΔE_g the change of grid voltage at the modulation frequency, reference to Fig. 2 gives:

$$\Delta E_g = E_r' \frac{Z'}{Z'+R_g}$$

The change of grid potential will be largest when the grid-leak condenser impedance Z' is great, but in no case can it exceed the rectified grid voltage E_r'.

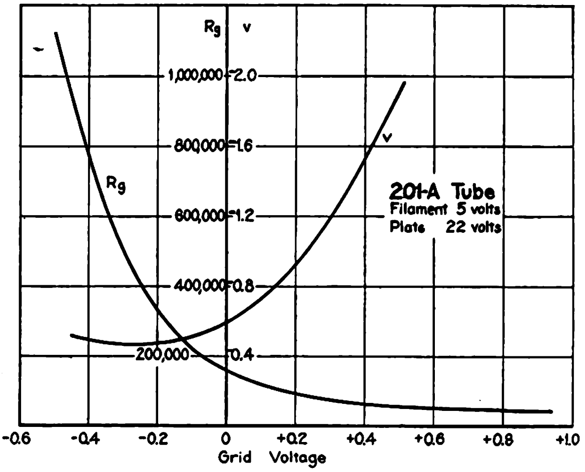

Fig. 3—Dynamic Grid Resistance and Voltage Constant As A Function of Grid Voltage.

The effect upon the plate circuit of the change of grid potential of the tube that takes place during detection can be determined by ordinary amplifier theory. In brief, a change of grid potential of ΔE_g is equivalent to inserting a voltage $-\mu\Delta E_g$ between the plate and filament of the tube in series with the dynamic plate resistance.

All equations essential to the computation of detector performance are given in Table I. Appendix I gives a complete set of computations for an actual numerical case.

DETECTION CONSTANTS

To set up the equivalent circuit of the vacuum-tube detector shown in Fig. 2 a knowledge of the dynamic grid resistance R_g is required. This is the resistance offered by the grid circuit of the detector tube to an increase of current caused by a slight increase in grid voltage. It is completely analogous to the dynamic plate resistance R_p, and depends upon the tube construction, and on plate, grid, and filament voltages.

The grid resistance of a typical vacuum tube is given in Fig. 3. The resistance, which is very low for positive values of grid potential, rises rapidly to high values when the grid is made negative by only a few tenths of a volt.

The quantity v that appears in Table I is the factor that takes into account the influence of tube characteristics upon the magnitude of rectified voltage produced when a signal is applied to the grid of the detector. This constant v has the dimension of a voltage, so can be appropriately called the *detector voltage constant.* The numerical value of the voltage constant v depends upon the operating grid potential, the plate voltage, and tube construction. Fig. 3 gives the effect of operating grid potential upon the voltage constant v for a typical tube. The rectified grid voltage is inversely proportional to v, so that an operating grid potential giving a small numerical value of v is desirable.

The detector voltage constant is determined by the grid resistance at the operating grid potential, and upon the rate at which the grid resistance changes with change of grid voltage. The exact relationship is given by the formula

$$\text{Detector voltage constant} = v = \frac{2R_g}{\dfrac{dR_g}{dE_g}}$$

Proof of this equation is given by Carson.[3]

A knowledge of the two detector constants, grid resistance R_g and voltage constant v, is all that is needed to compute the detector performance (i.e., to compute the change of grid voltage caused by the application of the signal voltage to the detector). Both of these constants can be measured by an alternating-current bridge suitable for measuring high resistances, such as shown in Fig. 4, in which the X arm of the bridge is supplied by the grid-filament resistance R_g.

The capacity C_1 balances out the grid-filament tube capacity and also stray capacities to ground of the 1000-cycle source. The input voltage to the bridge is controlled by resistance R_5 and current read by thermocouple milliammeter A. The grid potential is controlled by variable voltages supplied from potentiometers P_1 and P_2 and read by voltmeters V_1 and V_2. Potentiometer P_2 supplies small but accurately known increments to the grid potential furnished by P_1, and can be cut out by switch S_3 when these increments are not needed.

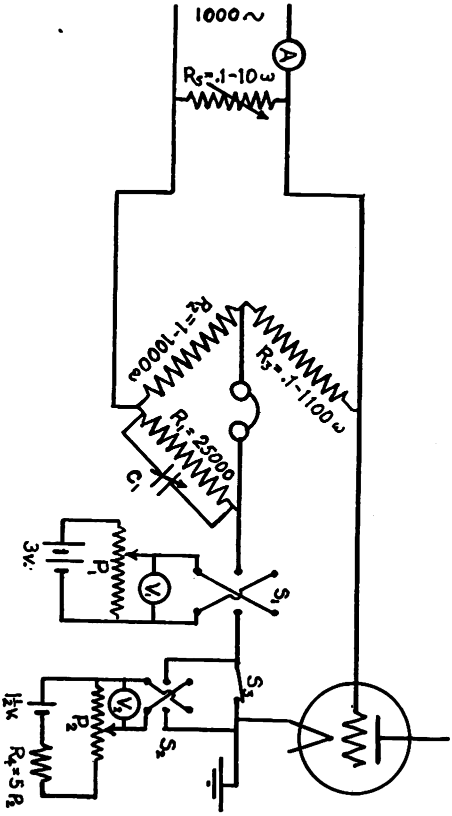

Fig. 4—Bridge Circuit for Measuring Detection Constants R_g and v.

Numerical values that have been found satisfactory for this equipment are given in Fig. 4.

To measure grid resistance with this equipment, potentiometer P_2 is cut out of the circuit by S_3, and the desired grid voltage is obtained by adjusting P_1 and the reversing switch S, The bridge is then balanced, which gives the grid resistance as

$$R_g = \frac{R_3}{R_2} R_1$$

The setting of the condenser C_1 does not enter into this result, although the condenser is necessary to obtain a null point.

The value of voltage constant v at this grid potential is obtained by measuring the grid resistance when the grid potential is higher and then lower than the potential at which v is desired, by a small amount ΔE. This increment of voltage can be obtained by connecting P_2 in series with P_1 by opening S_3. The increment ΔE is read by V_2 and can be made either positive or negative by the switch S_2.

Call E_0 the voltage read by V_1 (the potential at which R_g and v are desired), and let R_g' be the grid resistance at a grid potential of $(E_0+\Delta E)$, and R_g'' is the grid resistance at a grid potential $(E_0-\Delta E)$, then the grid resistance has been changed an amount $R_g'-R_g''$ by changing the grid voltage $2\Delta E$ volts. Accordingly

$$\text{Detector voltage constant} = V = \frac{2R_g}{\dfrac{(R_g'-R_g'')}{2\Delta E}}$$

The voltage constant is usually negative.

In measuring v and R_g, the input to the bridge should be kept low, preferably not over 0.10 volts effective. The voltage ΔE should also be small. Values of 0.05 volts have been found satisfactory. The sensitivity of the bridge can be increased by using an amplifier to detect balance.

This method of measuring R_g and v is quite accurate. The dynamic grid resistance can be measured to a number of significant figures that gives the difference $R_g'-R_g''$ with fair accuracy, and the increment of voltage ΔE is read directly on the millivoltmeter V_2.

THE GRID LEAK AND GRID CONDENSER

The primary function of the grid leak is to fix the operating grid potential, which is the potential of the grid return lead minus the voltage drop produced in the grid leak by the grid current, at a value most suitable for detection. The higher the grid resistance the more negative will be the actual grid potential.

The proper size grid condenser is a compromise between two conflicting requirements. It should be large to by-pass the radio-frequency signal voltage to the detector grid with a minimum of voltage drop, and it should be small in order to offer a high impedance to the rectified grid voltage.

The loss of signal voltage due to a finite grid condenser of capacity C is easily analyzed. Referring to Fig. 1, let E_{osc} be the signal voltage applied to the grid circuit, with E_s the part of this voltage that actually reaches the grid. Then[4]

$$E_s = E_{osc}\,\frac{C}{C + C_g}$$

where C_g is the effective grid-filament capacity of the tube to the radio-frequency signal, and is generally less than the corresponding value at audio frequencies.

It is important that the loss of signal voltage in the grid condenser be kept small as the rectified grid voltage is proportional to the square of the voltage applied to the grid. Using a large grid condenser to accomplish this gives a low grid-leak condenser impedance, however, and hence a poor utilization of the rectified grid voltage present. The best compromise seems to be a grid condenser about ten times the effective grid-filament capacity at radio frequencies. This leads to grid condensers of 150 to 250 $\mu\mu$f.

DETECTION OF RADIO-TELEPHONE AND TELEGRAPH SIGNALS

In the reception of telephone signals the rectified grid current contains the frequencies present in speech and music, and high-quality reception can be obtained only if all frequencies up to about 5000 cycles are about equally well reproduced.

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[4] This formula neglects the effect of the grid-filament resistance that in reality shunts the capacity C_g. When the grid condenser is large compared with C_g, as is always the case, this approximate formula is sufficiently exact for ordinary work, even at low radio frequencies where the error is greatest. The exact formula is

$$E_s = E_{osc}\,\frac{Z_g}{Z_g + 1/jwC}$$

where

$$Z_g = \frac{1}{1/R_g + jwC_g}$$

The most satisfactory way of obtaining equally good detection for all audio frequencies is to select an operating grid potential that makes the grid resistance R_g lower than the grid-leak condenser impedance at all audio frequencies to be preserved. Under this condition the change of grid potential is only slightly less than the rectified voltage E_r acting in the equivalent detector circuit, and no matter how much the grid-leak condenser impedance may change with frequency, this impedance, being always greater than R_g, will receive most of the rectified grid voltage across its terminals.

As the proper grid condenser capacity has already been fixed at about ten times the effective grid-filament capacity, choice of the most suitable value of leak resistance must now be made. For the reception of telephone signals this selection should be such as to give an operating grid potential that meets the following conditions as far as possible: (1) the grid resistance should be high in order to reduce the radio-frequency energy consumption by the grid circuit; (2) the grid resistance should not be so high as to cause the higher audio frequencies to be poorly detected; (3) the detector voltage constant should be as small as possible, in order that a given signal will generate a large rectified voltage, thus giving high sensitivity.

The operating grid potential which best meets these requirements is a potential which makes the dynamic grid resistance approximately equal to the reactance of the effective grid-condenser capacity at the highest audio frequency that is to be preserved in the detector output. This arrangement reproduces the highest audio frequency desired about 70 per cent as well as the lower notes, because at all but the lowest audio frequencies the grid-leak condenser impedance is substantially the impedance of the effective grid capacity taken alone. A lower grid resistance gives unnecessary energy losses in the grid circuit, and in most instances a larger value of v, both of which result in lowered sensitivity. Higher values of grid resistance than that suggested give rise to undue distortion of the high notes.

In receiving radio-telegraph signals the important component of the rectified grid voltage is the audio-frequency note heard in the phones, and is usually about 1000 cycles. The detector is accordingly operated to give the maximum possible change of grid potential at this frequency, without regard to the detection of other frequencies, which normally means a more negative grid (and hence higher leak resistance) than is the case with phone reception.

The effect of the operating grid potential (and hence of grid-leak resistance) upon the quality of detection is shown in Fig. 5, which has been computed for the tube characteristics given in Fig. 3. It is seen that less negative operating grid potentials obtained by the use of lower resistance grid leaks) give uniform detection for a wider range of frequencies than the more negative operating grid potentials (obtained by the use of high resistance grid leaks). A high grid-leak resistance gives better detection at low audio frequencies because the more negative operating grid potential gives a larger rectified voltage, but this gain is at the expense of poorer quality.

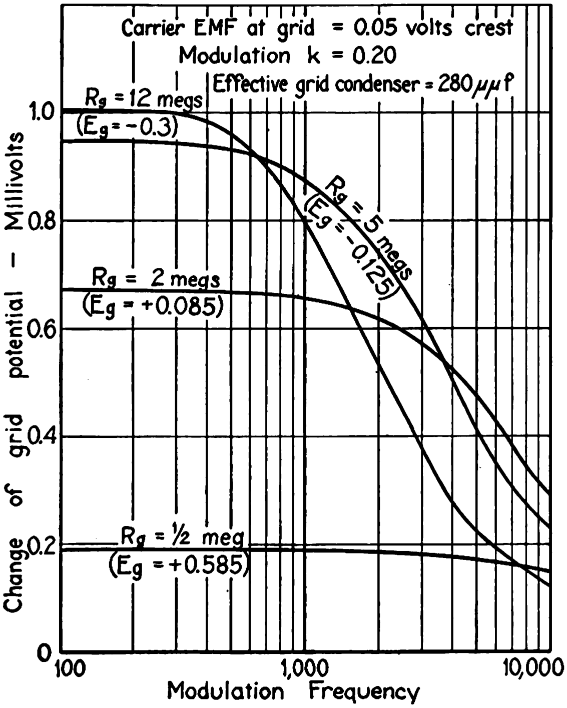

Fig. 5—Effect of Grid-Leak Resistance and Hence of Operating Grid Potential on the Performance of Detector of Fig. 2, Showing Crest Grid Potential Variation at Modulation Frequency As A Function of Modulation Frequency.

ASSUMPTIONS UNDERLYING THE THEORY OF DETECTORS GIVEN IN PAPER

The quantitative aspects of detector theory presented in this paper are subject to certain limitations. The fundamental assumptions are similar in nature to the assumptions underlying the equivalent circuit of the vacuum-tube amplifier, and are of the same order of magnitude.

The theory of detection as given holds exactly for infinitesimally small signal voltages, and its use also leads to correct results for all signal amplitudes which do not exceed the limits within which the detector current output is proportional to the square of the applied signal voltage. With high-vacuum tubes signal voltages of 0.05 volt effective and even higher come within the limit.

In the establishment of the equivalent detector circuit there is implied the assumption that the change of plate-filament voltage caused by the detection process has no effect on the action. Experiments have shown that this factor introduces no measurable error.

The treatment of detection that has been given applies to gaseous as well as high-vacuum tubes provided one works with signal amplitudes, values of ΔE, and bridge input voltages that are within the square-law limit of the tube characteristic. This requirement is easily met with hard tubes, but in gaseous tubes with "kinks" the limits are very small.

All plate-circuit rectification has been neglected. When present, such action can be computed separately, and the effects can be added to those produced by the grid action, taking into account the phase relations, which are normally more or less in opposition.

The analysis that has been made of detectors applies only to steady state conditions. The time constant of the grid circuit is so small, however, that transient changes are complete in less than the length of one cycle of the highest frequency that the detector satisfactorily reproduces.

Appendix I

NUMERICAL EXAMPLE

A radio-frequency voltage of 0.05 volts crest value modulated 20 per cent at 1000 cycles is applied to the grid condenser of the tube having the characteristics of Fig. 3, and operated at a grid potential of −0.125 volt. Essential information that would be known is:

R (giving desired grid potential)	5.0 megohms
R_g (from Fig. 3)	243,000 ohms
v (from Fig. 3)	−0.50 volts
Grid-filament tube capacity	
At radio frequencies	20$\mu\mu$f
At 1000 cycles	80$\mu\mu$f
Grid condenser	200$\mu\mu$f

The signal voltage actually applied to the grid is 0.05 (200/220) = 0.0454 volts. From Table I this voltage produces the following components of rectified grid voltage:

Direct-current component = 2.11 millivolts
1000-cycle component = 0.828 millivolts crest value
2000-cycle component = 0.0414 millivolts crest value

The 1000-cycle change of grid potential is found as follows. At 1000 cycles the impedance of the 5-megohm grid leak shunted by the effective grid condenser capacity of 280$\mu\mu$f is $565{,}000_{-83\text{ deg. }31\text{ min.}}$ ohms, and the impedance of the entire grid circuit is $565{,}000_{-83\text{ deg. }31\text{ min.}} + 243{,}000 = 644{,}000_{-61\text{ deg. }19\text{ min.}}$ ohms. The 1000-cycle change of grid voltage therefore has a crest value of 0.828(565000/644000) = 0.725 millivolts.

In a similar manner the direct-current component of change of grid potential is 2.11(5000000/5243000) = 2.01 millivolts and the 2000-cycle change of grid potential is 0.0414(284000/383000) = 0.0307 millivolts crest value.

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SIMPLE INDUCTANCE FORMULAS FOR RADIO COILS*

By
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IT is the purpose of this brief paper to present two simple formulas which the writer has found very useful for computing the inductance of the simple types of radio-frequency coils.

The new formulas are patterned after an empirical formula derived by Professor L. A. Hazeltine some years ago, for the inductance of a multi-layer coil. This formula follows, for dimensions in inches, according to Fig. 1:

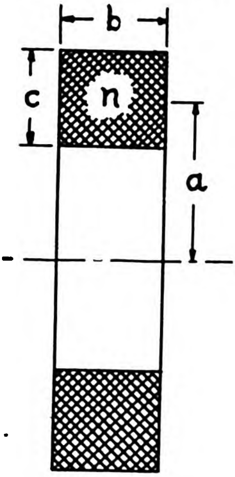

Fig. 1

$$L=\frac{0.8a^2n^2}{6a+9b+10c}\text{ microhenries} \tag{1}$$

This formula is correct to within about 1 per cent when the coil has approximately the shape shown in Fig. 1, such that the three terms in the denominator are about equal.

The new formulas were derived empirically from the inductance formulas and curves in Circular 74 of the Bureau of Standards. The corresponding coil formulas of this circular, however, either rely on tables or include expressions which are inconvenient to compute. For this reason there was a need for more convenient

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* Original Manuscript Received by the Institute, June 18, 1928.

formulas, even with the loss of some accuracy, for use in the laboratory. The formulas to be given are easy to remember and can usually be computed with one setting of the slide rule.

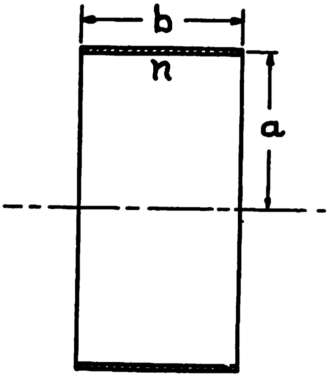

Fig. 2

One formula (derived in August, 1925) gives the inductance of a single-layer helical coil, in terms of the dimensions in inches, according to Fig. 2:

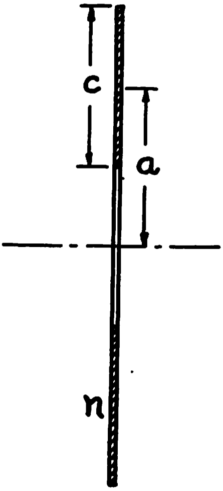

Fig. 3

$$L = \frac{a^2 n^2}{9a + 10b} \text{ microhenries} \tag{2}$$

This formula is correct to within 1 per cent for coils with $b > 0.8\,a$.

The other formula (derived in December, 1927) gives the inductance of a single-layer spiral (or helical) coil, in terms of the dimensions in inches, according to Fig. 3 (or Fig. 2):

$$L=\frac{a^2n^2}{8a+11c}\left(\text{ or }\frac{a^2n^2}{8a+11b}\right)\text{microhenries} \tag{3}$$

This formula is correct to within 5 per cent for coils with $c>0.2\,a$. (or $2a>b>0.2\,a$).

In general, formula (2) should be used for helical coils, whenever $b>a$, since (2) is more accurate under this condition. For shorter helical coils, however, formula (3)—in parentheses—is more accurate, when $a>b>0.2a$. The two formulas (2) and (3) give the same value when $b=a$.

In both formulas (2) and (3), the accuracy may be less than the stated accuracy when there are too few turns, when the spacing between turns is too great, or when the skin effect and distributed capacity are important factors.

Volume 16, Number 10 *October, 1928*

A GANG CAPACITOR TESTING DEVICE*

By

Virgil M. Graham

(Stromberg-Carlson Telephone Manufacturing Company, Rochester, New York.)

Summary—*Testing equipment for gang capacitors for factory inspection must be simple, accurate, and reliable in operation. The test described uses two oscillating circuits, one of which is tuned by each unit of the capacitor gang in succession. The arrangement is such that the first unit acts as a standard for the other units. The zero beat method of measurement is used. The other oscillator is tuned separately and acts as a reference. The complete procedure of operation is described.*

THE gang variable capacitors of from three to five units that are used in modern single selector radio receivers offer a considerable problem when they must be tested in production or inspected in large quantities. Inasmuch as the successful operation of the receiver depends on the accuracy of this gang capacitor these tests are of utmost importance. They must be accurate and reliable in operation and for factory use must be simple. There must not be difficult initial adjustments which are liable to become disturbed when used a short time without the operator knowing that anything is wrong.

In view of all this a gang capacitor test has been devised which meets the major requirements. Two oscillators are used, the frequency of the first being controlled in three steps with a vernier adjustment, and the frequency of the second being controlled by the units of the gang capacitor which are inserted successively in the circuit.

This test is designed to read the variation of capacitance of the units of the gang with respect to the first unit as a standard. This is more important than the absolute capacitance of each unit, as the average capacitance may vary over a much larger tolerance than can the capacitance of the units with respect to each other. The gang capacitor units are tested at three positions, 0 deg., 90 deg., and 180 deg. In each position the first unit acts as a standard. The capacitors that this test was designed for are of such design that these three test positions are sufficient to insure accuracy throughout the range. Of course, if necessary more test positions could be provided.

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* Original Manuscript Received by the Institute, June 28, 1928.

The first oscillator has three frequency settings, as stated above, corresponding to the three test positions of the gang capacitor. The frequency of this oscillator is three-fifths of that of the second oscillator at corresponding positions so that a zero beat can be obtained between the third harmonic of one and fifth harmonic of the other. This prevents the "pulling-in" effect. A vernier is provided in the first oscillator to compensate for variations in average capacitance of the gang capacitor.

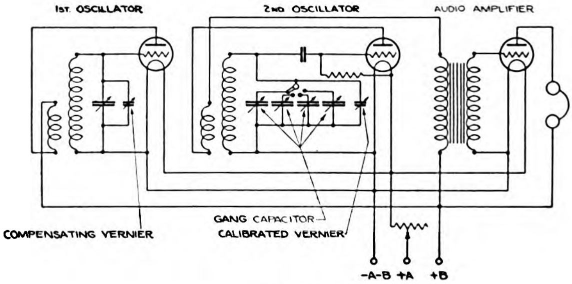

Fig. 1

The variation between units is indicated on a vernier capacitor which is connected across the second oscillator, the one in which each unit is connected successively. This vernier is calibrated in micro-microfarads.

The operation is as follows: The gang capacitor is inserted in the fixture, set at minimum (0 deg.), and the first unit is connected to the second oscillator, the calibrated vernier capacitor being set at zero. A "zero center" is used so that both plus and minus tolerances can be obtained. Next the first oscillator is set at the proper position and its vernier is adjusted so that a zero beat between the two oscillators is obtained. Now the first capacitor unit is disconnected from the second oscillator circuit and the second unit connected in. The calibrated vernier on this oscillator is adjusted until zero beat is again obtained. The reading of this vernier gives the difference in capacity between the first and second units of the gang capacitor. Of course, the first oscillator is not touched during this adjustment. The rest of the units of the gang are now connected successively and their capacitance with respect to the first unit obtained.

Now the gang capacitor is set at 90 deg. and at maximum, going through the same procedure with each of these positions.

It can be readily seen that a slow drift in the oscillators due to battery voltage changing, etc., while not desirable would not be particularly serious as the first oscillator vernier is adjusted with every gang capacitor. This means that the reference is constantly checked and as long as there are no momentary changes the outfit cannot give trouble from poor adjustment.

Fig. 1 shows the circuit of the arrangement.

Volume 16, Number 10 *October, 1928*

RADIOTELEGRAPHIC CENTER AT ROME (SAN PAOLO)*

BY

G. PESSION AND G. MONTEFINALE
(Rome, Italy)

Summary—*The radiotelegraphic station at San Paolo constitutes*[1] *the first and largest continuous wave radiotelegraphic installation (Poulsen arc of 250 kilowatts) constructed in Italy in 1917.*

The service with the Poulsen arc, which has been completely satisfactory, reached its greatest development in 1922, in which year there were transmitted 2,175,000 paid words directly to stations in East Africa, United States, England, Germany, Switzerland, Norway, Roumania, and ships and stations in Egypt.

The decrease in the number of transmitted words in 1923 resulted from the repairing of the Poulsen arcs and the complete replacement of the towers, for which reasons the station of San Paolo was inactive for six months, during which time the service went through the new Coltano[2] *station temporarily.*

The complete reconstruction of the 218-meter wooden towers which had been standing for only six years was a necessary consequence of the rapid war-time construction for the purpose of establishing more direct communication with America and with the colonies. After repairing the towers it was possible to erect two or more of the same type with some of the remaining timber, still in good condition; one tower 124 meters in height, used as an anchor tower for the tail of the antenna, allowing less sag; the other 86 meters in height, used as a support for the auxiliary antennas. The reconstruction was quickly accomplished and at a low cost.

The old antenna, of the triangular net form made of square mesh, was replaced by a lighter one, of the horizontal fan type, of 16-strand phosphor bronze wires 4 millimeters in diameter. Later a multiple grounding system was constructed, using one telegraphic wire 3 millimeters in diameter.

The efficiency of the station was improved because it was possible to obtain in the antenna a current of 120 amperes for the 10,750-meter wave by applying to the arc a reduced voltage of 500 volts instead of 800 volts, which was formerly necessary.

Notwithstanding its good performance, the Poulsen arc station at San Paolo, which is located on the outskirts of Rome, caused a great deal of trouble

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* Original Manuscript Received by the Institute, January 4, 1928. Paper presented at the thirty-second annual meeting of the A.E.I., Como, Italy, October, 1927. Translated from *L'Elettrotecnica*, Journal of the Associazione Elettrotecnica Italiano, XIV, No. 19; July 5, 1927.

[1] B. Micchiardi, G. Pession, G. Vallauri, "The Radiotelegraphic Station at Rome, San Paolo." *L'Elettrotecnica*, Maggio 1920, Vol. VII, N. 13–14 e Pubblicazione, N. 8, dell' Istituto E. e R.T. della R. Marina. PROC. I.R.E., 8, p. 142; April, 1920.

[2] G. Vallauri, "The Radiotelegraphic Center at Coltano," *L'Elettrotecnica*, Gennaio 1924, Vol. IX, N. 1, e Pubblicazione, N. 28, dell' Istituto E. e R. T. della R. Marina. Refer also to earlier paper published in PROC. I. R. E.

to public and private services in the city and surrounding country. The characteristic noise and numerous harmonics brought on by the operation of the Poulsen arc are responsible for this.

In San Paolo, while the installation work was progressing on a new, powerful long-wave tube apparatus already designed to purify the harmonic waves of the Poulsen arc transmission by the addition of an adequate intermediate circuit, the short wave was utilized, which was a development of great importance rapidly accomplished in 1924. This change permitted the development of an entirely new method of transmission which attained the same results with less expensive apparatus. This apparatus has been fully described.[3,4]

I. SHORT-WAVE SERVICE OF THE SAN PAOLO RADIOTELEGRAPHIC STATION

IN the San Paolo station in 1926 besides the Poulsen arc transmitter of 250 kilowatts (10,750-meter wave) the following were in service:

(a) One 6-kilowatt tube transmitter operating on 66-meter wavelength.

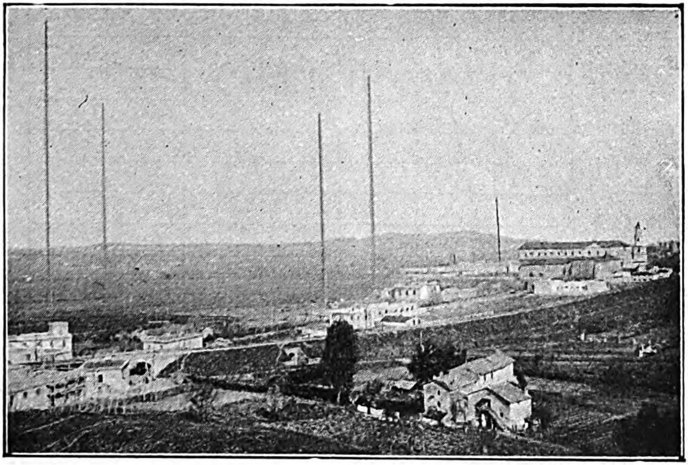

Fig. 1—View of the San Paolo Radiotelegraphic Station.

(b) One 6-kilowatt transmitting tube set operating on 32-meter wavelength.

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[3] G. Pession, A. Pizzuti, "Experimental Contribution on Short-Wave Radiotelegraphy," *L'Elettrotecnica*, Marzo 1925, Vol. XII. N. 7 e Pubblicazione N. 32 dell' Istituto E. e R. T. della R. Marina.

[4] G. Pession, G. Montefinale, "Commercial Radiotelegraphic Experiments with Short Waves." *L'Elettrotecnica*, 15 Luglio 1926, Vol. XIII, N. 20 e Pubblicazione N. 36 dell' Istituto E. e R. T. della R. Marina.

II. PERFECTING THE RECEIVING STATION

Notable improvements were made in the Monterotondo receiving station, which made possible the reception of long-waves of 600 meters and up by the use of an external receiving coil, and an ordinary antenna supported by two trees 32 meters high for the reception of intermediate waves (2000 meters) from ships. There were also specially designed antennas for short-wave reception. (Beverage).[5]

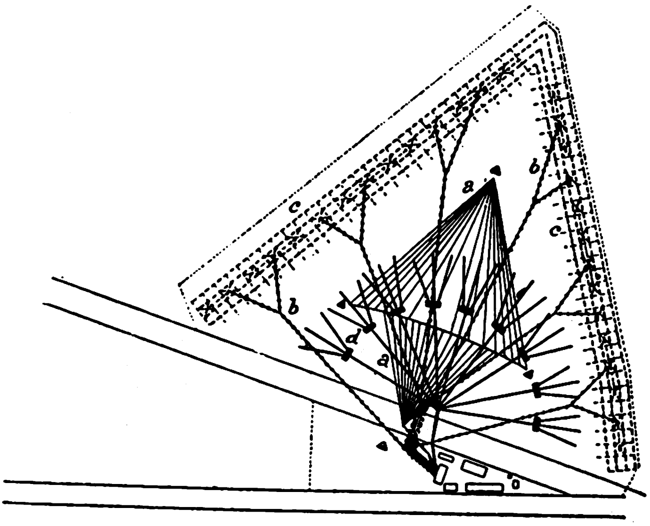

Fig. 3

It was possible to provide for the reception of signals in various wave bands as follows:

(a) Long-wave reception by means of outside loop antenna.
(b) Intermediate wave reception with ordinary antenna.
(c) Far East (Pekin) reception with 57-meter wavelength.
(d) Italian Somaliland (Afgoi) reception with 38- and 70-meter wavelength.
(e) Red Sea Colonies reception with 60-meter wavelength.
(f) Near East (Rodi) reception with 73-meter wavelength.
(g) Cyrenaica (Bengazi) reception with 53-meter wavelength.
(h) Tripolitania (Tripoli) reception with 51-meter wavelength.

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[5] G. Pession, G. Montefinale, loc. cit.

The (a) service is of a sporadic character, and (b) continues day and night. All other services work on fixed time schedules.

Normally, the services (b), (c), and (d) are carried simultaneously from 4:00 P.M. to 2:00 A.M. G. M. T., which is the best time for the station to operate.

The short-wave reception for the three transcontinental stations, Pekin, Afgoi, and Asmara, and from the sea stations on the Mediterranean at Rodi, Bengazi, and Tripoli was had with great regularity under all atmospheric conditions, even during electrical storms, which are very frequent. During 1926 and also during the first months of 1927, there has not been a single service interruption, as contrasted with the many difficulties with long-wave reception which characterized the preceding years.

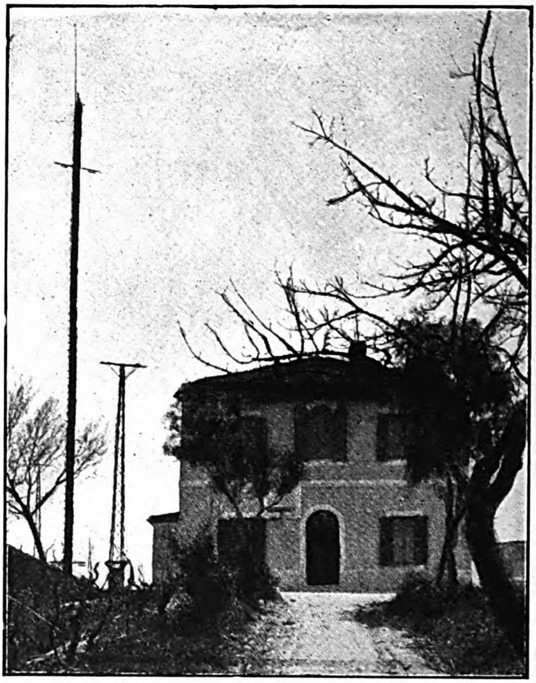

Fig. 4—Outside View of the Monterotondo Receiving Station.

As an example, the monthly reports of words received by the short-wave receiving set at the Monterotondo station can be consulted.

TABLE II

SUMMARY OF THE RECEPTION WORK ON SHORT WAVE; COPIED FROM MONTEROTONDO STATION, APRIL, 1927

Station of Origin	Number of Received Telegrams	Number of Received Words
Ships	101	2,705
Pekin	617	26,447
Afgoi	2,334	52,023
Asmara	2,284	55,832
Rodi	2,405	43,882
Bengazi	258	4,222
Tripoli	90	3,463
Total	8,089	188,574

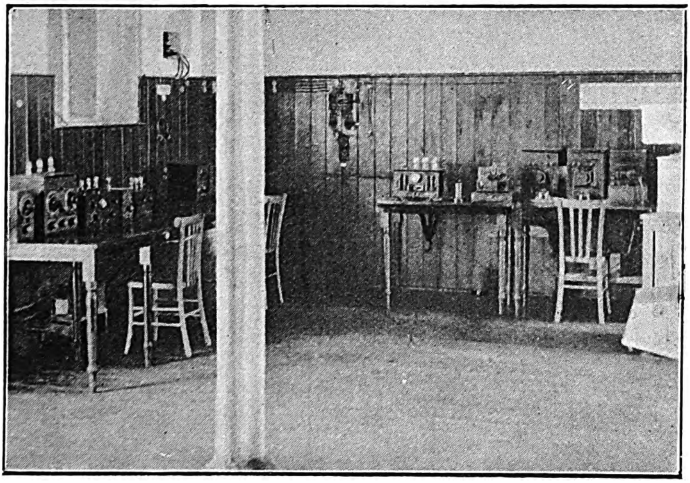

Fig. 5—Interior of the Monterotondo Receiving Station.

III. CAPACITY INCREASE OF THE SHORT-WAVE SETS

Any station similar to that of San Paolo must be able to transmit simultaneously to several corresponding stations; also to ships on the high seas at any time, or at least during the main part of the day, and to the greatest possible distance.

Experience has demonstrated that the arc installation of the San Paolo station has a range not greater than 4500 kilometers. The small 6-kilowatt short-wave stations permit better communication to a greater distance, but only for a few hours.

Therefore the problem of increasing the power of transmission arose immediately. It seemed logical to discontinue the use of long-wave sets and to substitute relatively low-power, short-wave sets at the station. As a solution of the problem the increase of the transmission power by using waves of the same order of length was therefore considered practical.

It was then decided to experiment with new types of tubes with water-cooled plates, which had the advantage of operation at relatively high power with such units connected in parallel. This would reduce the effect of the capacity between electrodes in oscillatory circuits, which is of particular importance in short and ultra short-wave work.

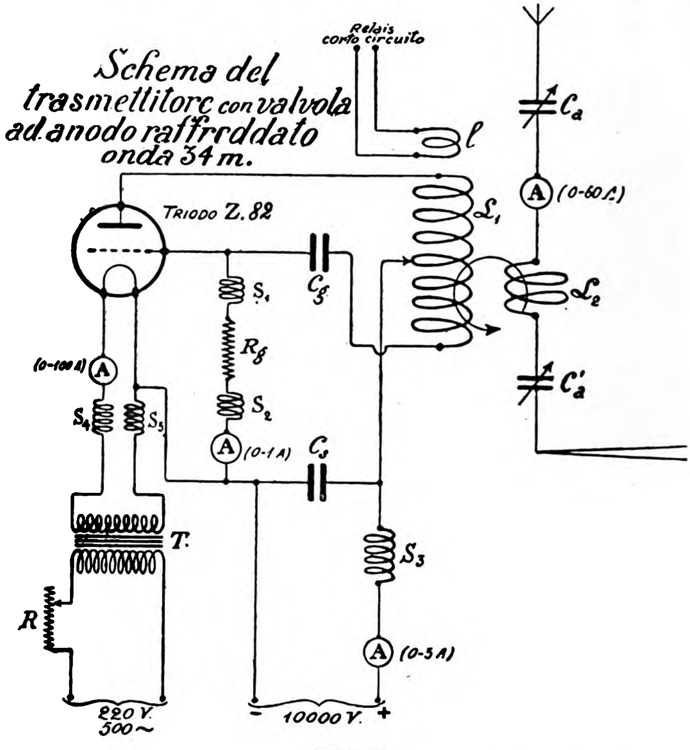

Fig. 6

The selection of tubes after the experiments favored the Ditta Philips 25-kilowatt Z.82 tubes for the production of short waves.[6]

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[6] The general layout and operation of the three-electrode tube and reservoir are shown in Fig. 7. The main data for its operation are:

Voltage for Filament Lighting	16–17 volts
Current for Filament Lighting	17 amperes
Plate Voltage	about 12,000
Saturation Current	near 8 amperes
Input	25 kw
Maximum Diameter	160 mm
Maximum Length	700 mm
Internal Capacity, Plate-Grid	29 $\mu\mu$f
Internal Capacity, Plate-Filament	11 $\mu\mu$f
Internal Capacity, Grid-Filament	3 $\mu\mu$f

The details of construction of the three-electrode Z. 82 Philips tube are described in the Ditta Publication. The plate is formed of nickel-

IV. 34-METER SHORT-WAVE GENERATING CIRCUIT WITH WATER-COOLED TUBES

For the construction of the new transmitter sets with selected 25-kilowatt water-cooled tubes, the circuit indicated in Fig. 6 was used for the 34-meter wave.

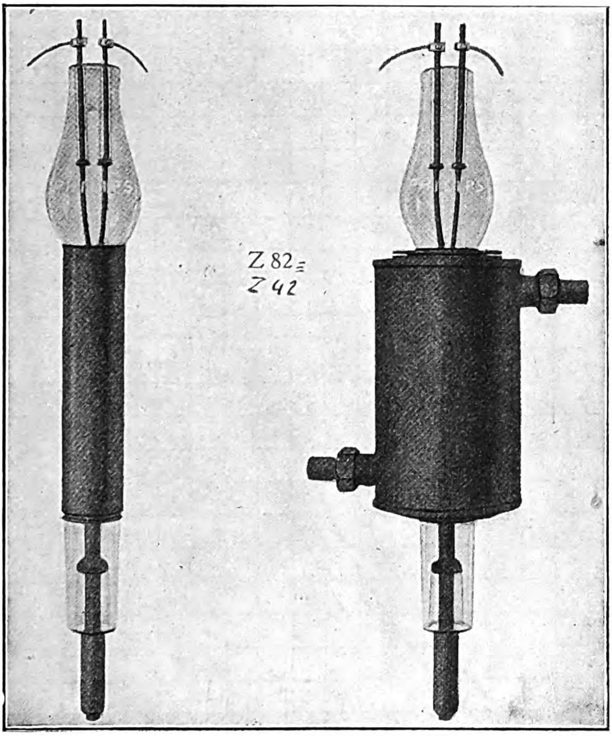

Fig. 7

The insulation of the various parts of the 25-kilowatt sets was taken into particular consideration, using special pyrex glass

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iron cylindrical sheath, welded at the ends by a special process and protected by two glass prolongations that, together with the plate, complete the whole enclosure of the three-electrode tube. The filament is made of 8/10 mm. diameter tungsten wires connected in parallel. The grid is made from a cylindrical metal mesh concentric with the filament. Both ends of the filament came out at the upper neck of the glass and those of the grid at the lower one. The grid is 22 cm. long, the plate 24 cm. and the filament about 18 cm. The diameter of the plate is 6 cm. Fig. 8 contains the plate characteristic curves of one water-cooled tube Z. 82 operated with negative grid potential of 400 volts and plate potential of 8,000, 10,000, and 12,000 volts.

insulators built according to Royal Navy design by M. I. V. A. Acqui Society (Italy).

In the regulation of the 34-meter short wave, it was found necessary to place the two impedances of the S_1 and S_2 coils on either side of the grid resistance, with the impedance S_4 and S_5 correctly proportioned at both ends of the filament. The keying was accomplished by using the absorption method (with counter wave), short circuiting the two turns of the connected inductance during the pause by means of specially designed relays.

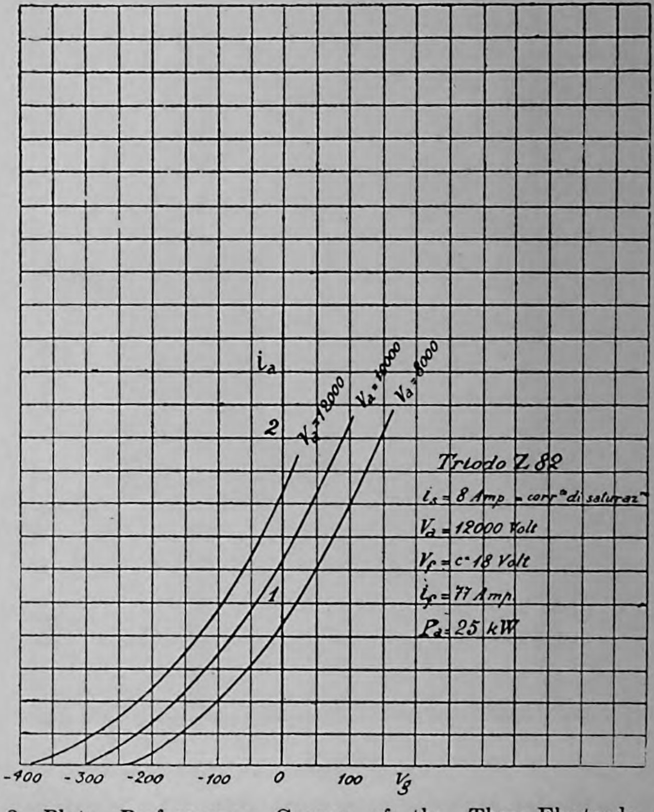

Fig. 8—Plate Performance Curves of the Three-Electrode Tube, Operated with Negative Grid Potential of 400 volts and Plate Potential of 8000, 10,000 and 12,000 volts.

The voltage necessary to light the filament of the 25-kilowatt three-electrode tube was furnished in exactly the same way as the other short-wave transmitter at the San Paolo station, reducing by means of a suitable transformer the 220-volt, 500-cycle current produced by a 10-kilowatt converter supplied by the 500-volt, 44-cycle city mains. The regulation of the filament lighting requires considerable precision and could be made either by varying the alternator excitation or by using a variable iron-core impedance coil of special design in series with the 220-volt primary circuit.

The plate voltage of 10,000 was obtained by rectifying the 500-cycle single-phase alternating current supplied by a 75-kilovolt-ampere alternator (Fig. 9). The pulsating current furnished by the rectification of the 500 cycles was smoothed by means of a filter made of impedance coils, 3 henries, each connected in series, and two C_1 and C_2 condenser banks of 1.3 μf each.

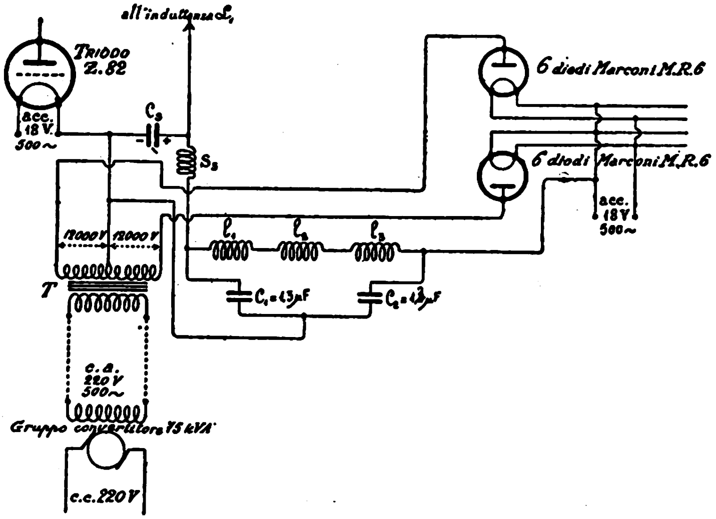

Fig. 9—Circuit of 220-volt, 500-cycle Rectifier with Six Marconi Two-Electrode Tubes in Each Side.

The external cooling of the plate by water circulation was arranged by a connection to the pipe which delivered the water for cooling the Poulsen arc. To obtain a convenient pressure, a tank was placed about 60 centimeters higher than the lower neck of the tube. From the tank a coiled rubber tube 12 millimeters in diameter and 10 meters in length (Fig. 10) delivered cold water to the lower orifice of the refrigerator. The water heated by the plate's dissipated energy is discharged from the upper orifice by means of another coiled rubber tube of the same length. Moreover, an efficient system of cooling was provided for the plate. Compressed air at a pressure of not less than 60

8 (very strong) were received. A more complete test was made on December 7, 1926, transmitting during the first 20 minutes of every hour from 12 A.M. to 6 P.M. G. M. T. The use of the 6-kilowatt set did not provide short-wave service with distant stations.

From such tests it can be concluded that by employing 34-meter length waves with 15-kilowatt power, it was possible to give a 12 hours' continuous service during the winter season with the Far East, an 18 hours' service with Somaliland, 20 hours' service with Eritrea, and practically continuous service with the station of the Dodecaneso and North African colonies.

It was apparent from the tests that, notwithstanding the increase of power, it was impossible to eliminate the silent zones at short distances from the station, which increased at night and gradually decreased as daylight came on.

With the installation described, using 20-kilowatt power, it would be possible to have a practical daytime range of about 6000 kilometers, a distance greater than that obtained by using the old 250-kilowatt long-wave Poulsen arc.

A station identical to that of San Paolo was constructed at Spezia and exhibited at the Volta Exposition at Como.

VI. COMPLETION OF THE CENTRAL STATION AT SAN PAOLO IN 1927

The installation at San Paolo was completed with an intermediate wave transmitter of modern type, powerful enough to establish communication with the Mediterranean Sea stations.

A 15-kilowatt tube transmitter was constructed which consisted of 12 Marconi M. T. 6 three-electrode tubes and a large inductance for coupling to the antenna, as represented in Fig. 11. The filament lighting of the M. T. 6 bulbs was made similar to that of other tube sets installed at San Paolo with 500-cycle single-phase power. The plate potential of about 6000 volts needed to operate the tube could be obtained either from the rectifier which supplied the plate potential to the 6-kilowatt short-wave set, or from the 500-cycle converter (Fig. 10) that feeds the short-wave station using the 25-kilowatt metallic tubes.

The high-frequency current generated by the set mentioned above was fed to the big antenna at San Paolo which, because of its great power of radiation and its low ground resistance, offered particularly favorable conditions. With the use of such an antenna it was easily possible to reach the 4800-meter service wave, the current at the base of the antenna being about 38 amperes. By inserting in the antenna an air condenser made by two condensers connected in parallel (each containing 26 cm by 60 cm plates separated 45 mm), a 2250-meter wave with a current of about 18 amperes was obtained.

By the use of the method just described (i.e., for 2250- and 4800-meter waves) the San Paolo station could communicate easily with every European station and any point on the Mediterranean Sea. There was also installed at San Paolo a special short-wave auxiliary set devised for those communications in which the 32-meter (6-kilowatt) short-wave transmission was not heard (silent zone).

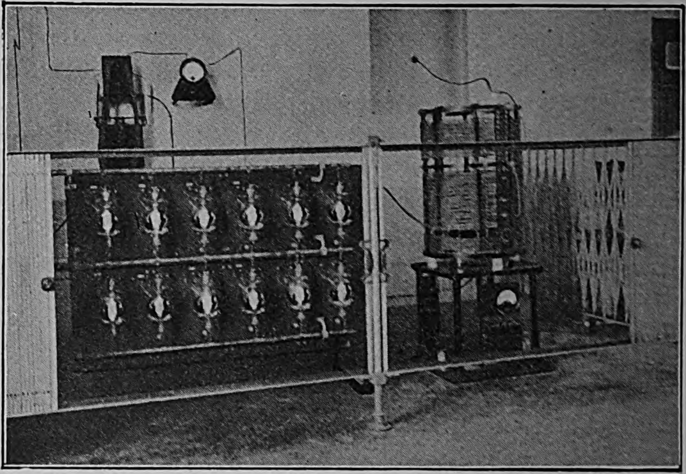

Fig. 11—15-Kilowatt Long-Wave Tube Transmitting Set.

The 3-kilowatt set was constructed entirely at the San Paolo Laboratory using the same diagram adapted for the transmitter of 25-kilowatt cooled electrodes, employing as an oscillator three Marconi M. T. 6 three-electrode tubes in parallel, and rectifying the 44-cycle public service power. This rectifier was made up of M. R. 6 two-electrode tubes, an equalizing circuit of 3 μf condensers, and a 20-henry impedance coil.

The peculiarity of such a transmitting system is that it acts synchronously with the radiotelegraphic receiving station at Rome. Operation from this station is made by means of a special telegraphic line.

After investigation, a relay system was installed by means of which the operator in service at the receiving station could light the tube filaments, apply the plate potential, and start the blower for cooling. Likewise San Paolo could transmit with short waves, not only to remote stations, which was the principal aim, but also to ships on the Tyrrhenian Sea.

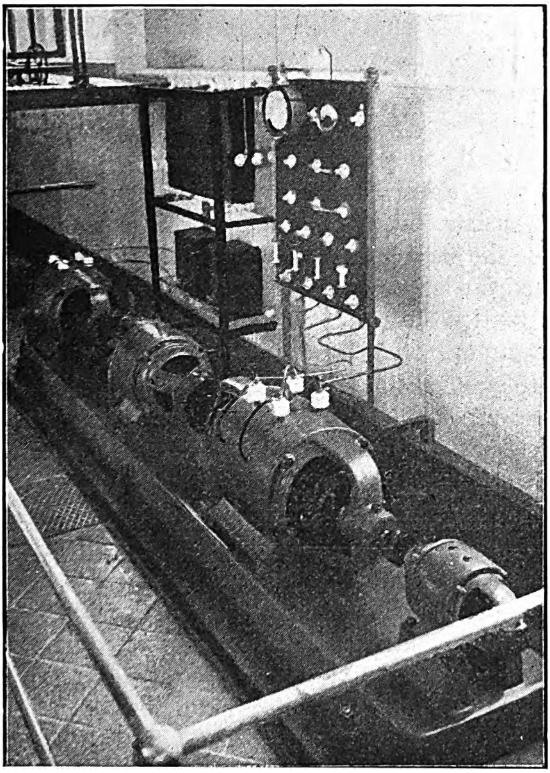

Fig. 12—High Tension Generating Unit (6000 volts) of the San Paolo Radiotelegraphic Station.

To make transmission tests using a more balanced plate voltage than was obtained by rectifying single-phase or three-phase alternating current, there was installed a high-tension generating set (Fig. 12) which furnished 6000 volts direct current necessary to operate the 66-meter wave apparatus.

This generating set built by the Ditta "Bayerische Elektricitats-Werke" is made up of a 6.9-kilowatt, three-phase, 500/290-volt, 44-cycle asynchronous motor with two 3-kilowatt direct-current generators driven directly coupled to both ends of the motor shaft, with 110-volt separate excitation, each one capable of furnishing 1 ampere and 3000 volts at 2550 r.p.m. Normally the generators are connected in series, thus obtaining the plate voltage of 6000 volts direct current.

The most important conclusion drawn from experience at San Paolo is the perfect regularity of the short wave. During service hours it was noted that the signals were of constant strength and nearly independent of atmospheric conditions. Fading occurred rarely and, in general, produced no appreciable disturbance to the service.

As a rule, all the service in connection with permanent stations and with foreign ships was performed in a satisfactory manner by using the 32-meter (6-kilowatt) short-wave transmitter while the service of the 34-meter (25-kilowatt) short wave was reserved for larger power requirements. During the winter the Pekin station received the 66-meter wave better, whereas during the summer the 32-meter was preferable.

From the numerous tests made, the behavior of the different waves used at San Paolo was perfectly understood. From 106 meters down to 32 meters (wavelengths of 106 meters, 80 meters, 66 meters, 50 meters, 40 meters, 34 meters, 32 meters were under experiment) it was clearly noticed that the longer wave shows the difference between day and night ranges better. With 106 meters and 80 meters, the day range is reduced much more than at night (about 25 times that of the day with devices used at the San Paolo station). The 50-meter wave gives a considerable day range. Going down to 40 meters, near which the silent zone begins to be noticed at night, the reception decreases rapidly even at distances of 50 to 100 kilometers, varying with the power used. After the reception has been secured in a more or less extensive zone, depending on the station and the hour, it comes back and is very intense even at a great distance.

The 32-meter wave has a day range much better than the preceding ones. At night the silent zone is remarkably complete at skip distances, ranging from 500 to 1500 kilometers, depending on the season. Using a 32-meter (6-kilowatt) wave, it is possible, for instance, to communicate during the daytime with ships on the North African and Egyptian coasts, whereas during the night communication is defective and at certain hours impossible with stations at Rodi. Frequently communication is not obtained at all even with stations at Spezia, Maddalena, Napoli, etc. On the other hand, signals from Mogadiscio, Massaua, and Pekin are very strong.

It has certainly been demonstrated that atmospheric disturbances decrease as the wavelength decreases. The following is an example: at one of the stations in Somaliland a 70-meter (6-kilowatt) set was operated together with a 38-meter set with smaller input. On many evenings reception at Monterotondo was impossible when using 70-meter wavelength on account of atmospheric disturbances, while it was easily obtained with 38-meter wavelength. For this reason the 38-meter wave was used to clear up the traffic.

Some stations in the Far East have reported satisfactory reception during the day for many hours, on a 15-meter wavelength.

As to the properties of the short waves over close ranges, the observations made at San Paolo correspond perfectly with those obtained at the Laboratory at Schenectady, with wavelengths of about 33, 65, and 109 meters.[7]

Some technical experiments that were conducted will be briefly described here. In particular:

1. To find the most convenient type of antenna for both transmission and reception.
2. To improve the stability of the frequency and the tone modulation.

Nothing positive was found for the transmitting antenna since none of the numerous experiments yielded results either better or worse than the rest.

It could be proved that there was a perfect equivalence between the vertical 30 deg. to 40 deg. angle type V antenna excited by a neighboring electrical field vibrating with a certain number of semi-waves (as are those previously described in other publications[8]) and the simple antenna slightly inclined, similar to that known as the Hertz type, the radiating part of which is formed of a single wire and a convenient Hertzian double wire feed at the center.

No noticeable difference was found in the intensity of the signals by varying the height of the suspension point successively. As an example there could be cited the experiment conducted September 13, 1926, when the antenna wires were extended horizontally 5 meters above the ground during transmission with a 66-meter wave. There was obtained in the stations in contact with other stations such as Pekin, Afgoi, and Chisimaio the same intensity of signals as that of stations transmitting with high antenna.

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[7] M. L. Prescott, "Tests of Radio Propagation on Short Wavelengths." *General Electric Review*, 30, No. 2; February, 1927.

[8] G. Pession, G. Montefinale, loc. cit.

Similarly, the commonly-used 0.011 μf fan-form antenna of 3500-meter natural wavelength was tested, transmitting with a short wave. Even with this the corresponding distant station did not show any difference in strength of signals in the telephone. Identical results have been found using simple, single-conductor antennas and the flat or tubular multiple conductor, which have been used normally in the service with intermediate and long-wave transmission.

In regard to the reception, it was proved conclusively that of all types of antennas described above, the best proved to be the horizontal one of very little elevation and grounded at the other end of the receiving set.

With this antenna, known as the Beverage antenna,[9] it is possible to obtain stronger signals than with the vertical one, especially at night, together with a remarkable immunity from atmospheric disturbances.

One night when some experiments on atmospheric disturbances were being carried out during a very violent storm with many electrical discharges that made impossible any services using long-wave transmission and necessitated grounding the telegraphic line, the best reception from Rodi, Asmara, and Afgoi was noted. Only occasionally some letters were lost in the noise of the storm that caused a deluge upon the station and short-circuited the receiver.

After testing various types of receiving sets it was found that the superheterodyne circuit was the best. For practical purposes, however, a simplified receiver with a regenerative tube in series with an audio-frequency stage through an audio-frequency filter and two stages of amplification was preferred.

The signals were very strong and constant and their intensity could be well regulated upon the filter coupler by a very loose coupling. The received voltage was sufficient to override atmospheric disturbances. Small variations in frequency were easily followed without causing the operator to lose the signals. This set utilized a Wheatstone Recorder.

[9] G. Pession, G. Montefinale, loc. cit.

Tests were conducted to determine the shortest distance to which the transmission of duplex service was possible with certain wave differences. With the filter, this distance could be reduced to a minimum, using a slightly different wavelength. Actually, it was fixed at 500 meters as the minimum practical distance with which the San Paolo station receiver could operate, when the 34-meter transmitter was being used. For other lengths, as 38-meter or more, this distance became greater.

In regard to tests made to improve the stability of frequency, the following results were obtained:

With the simple diagrams and schemes used, it was impossible to obtain a pure and constant note, although well-rectified direct current was used to supply the plate. This failure was blamed on the use of alternating current for lighting the filament, but even more on the irregularity of tube operation. The capacity of the tube is an essential element in determining wavelength. This capacity undergoes certain variations during operation caused either by the unavoidable small vibration or by the capacity being shunted by the internal resistance of the tube. The latter is not perfectly constant. The note became worse with the increase of power as well as with the shortening of the wavelength. The tone of transmission in the San Paolo station was sufficiently constant and clear to permit good reception. It was considered unnecessary to use a more complicated diagram and scheme (oscillatory excitation and crystals) which would be justified only in the case of rapid and commercial service. For practical purposes, the simple sets of the San Paolo station had been demonstrated to be sufficient for the service required, when conducted by skilled personnel.[10]

The variations of frequency over rather long periods of time, which before were serious, proved to be greatly dependent on the variations of the voltage applied to the filament. To eliminate this, a separate voltage regulator was used for each set.

The remarkable improvements made in the radiotelegraphic center of San Paolo, Rome, during 1926 and 1927 were particularly accomplished by Lieutenant Baccarani Ugo, the radiotelegraph operator in charge of the San Paolo station, and Antonelli Fabrizio, who constructed the sets and conducted the test and experiments.

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[10] In the course of 1928 the long-wave transmission and reception at San Paolo has practically ceased. All the traffic of 400,000 monthly paid words is developed only with short-wave sets.

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Discussion on

ON THE DISTORTIONLESS RECEPTION OF A MODULATED WAVE AND ITS RELATION TO SELECTIVITY*

(F. K. VREELAND)

R. Raven-Hart†: Dr. Vreeland does not appear to have taken into consideration the "demodulating" effect of a stronger on a weaker signal, to use the term suggested by Dr. Beatty[1].

If the condition stated by Dr. Beatty, that the carrier wave of the weaker signal must be reduced to one-tenth that of the stronger one, is valid (and personal observations would lead me to believe that even this condition is too stringent, at least for French tubes: it depends of course on the degree of linearity of the detector characteristic), then the selectivity resultant on reducing the interfering signal strength to 1.5 per cent of the signal carrier frequency cannot be considered as merely "passable."

In fact, sufficient selectivity would be obtained with a curve intermeditate to those for effective resistances of 1 and 2 per cent on Dr. Vreeland's Fig. 3, and such a curve would give a response at 10 kilocycles distance of about 40 per cent of the carrier, assuming curves such as Dr. Vreeland shows.

If, however, curves such as those given by Mr. Jones[2] are assumed, this value of 40 per cent becomes more like 70 per cent which is probably acceptable, especially in view of the fact that these extreme frequencies are rarely transmitted in practice.

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* Presented at the Annual Convention of the Institute of Radio Engineers, January 9, 1928. PROC. I.R.E., **16**, 255; March, 1928.

† Consulting Engineer, Paris, France.

[1] "The Apparent Demodulation of a Weak Station by a Stronger One." *Wireless Engineer;* June, 1928.

[2] PROC. I.R.E., **16**, 671; May, 1928.

Volume 16, Number 10 *October, 1928*

A DECIMAL CLASSIFICATION OF RADIO SUBJECTS; AN EXTENSION OF THE DEWEY SYSTEM*

PREPARED BY
BUREAU OF STANDARDS, WASHINGTON, D. C.

IN 1923 the Bureau of Standards published Circular 138, "A Decimal Classification of Radio Subjects—An Extension of the Dewey System."[1] The reference numbers used in the monthly list of references to current radio literature now published in the PROCEEDINGS of the Institute of Radio Engineers are according to this classification. The Bureau of Standards would be glad to receive suggestions or criticisms on the following proposed revision of this circular.

I. Need for Classification

The radio laboratory of the Bureau of Standards has, in common with other workers in the radio field, felt the need for a systematic scheme of classification for subjects in radio science and engineering. This need has been felt not only for use in classifying the references to current radio publications but also for classifying other radio material, such as drawings, books, reports, etc. In an effort to fill the need for a radio classification the present extension of the Dewey decimal system has been prepared.

Such a system makes it easy to place books on related subjects near together on the shelves or to file references on the same subject all in the same group and not by the order of their addition to the collection or file. If a classification is to be of the most use any part of it must be capable of expansion, or it must be possible to disregard any part of the classification without interfering with the usefulness of the remaining parts. These requirements are met by the classification given below.

II. The Dewey Decimal System of Classification

Under the Dewey decimal system,[2] of which the present classification is an extension, classification is by subject, numbers

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* Original Manuscript Received by the Institute, August 24, 1928.

[1] Copies of this circular may be obtained from the Superintendent of Documents, Government Printing Office, Washington, D. C., price ten cents.

[2] M. Dewey. Decimal classification and relative index for libraries, clipping notes, etc. 10th Ed. Lake Placid Club, N. Y. Forrest Press, 1919.

being used to show the relative positions of the books, cards, or other material. The numbers, therefore, show both what the material is (that is, its subject matter) and where the material is (that is, its location on the shelves or in the files). In the classification list the indentation and the figures prefixed to each item show the rank of each subject in the classification.

Accompanying the classification is an index which is arranged in the usual alphabetical order. References are made in this index to the subject classification number rather than to pages or to arbitrary shelf numbers. The index is used in determining the number to assign to a given item or material or to learn where to place it in the files. The index is also used by any person desiring to locate the material covering a given subject. The reference number tells immediately where all material on that and on related subjects can be found.

a. OUTLINE OF CLASSIFICATION

The whole subject of radio is put in its proper place in the Dewey classification—621.384. The relation of this place to the general field is shown by the following table.

Class	600.	Useful arts
	20.	Engineering
	1.	Mechanical
	.300	Electrical
	.080	Communication
	.004	Radio

In a strictly radio library or office it is convenient to represent the figure 621.384 by "R" and this abbreviation is used below in the further classification of radio. Thus, R211 indicates 621.384.211.

While some of the details of the Dewey system itself seem at the present time to be illogical (for example, electrical engineering a subdivision of mechanical engineering), the system has been widely adopted, and confusion would result from attempts to change it into a more logical form. The Dewey system has some general features which are found especially advantageous. For example, all general material under a given class is put under the class itself (usually having a final figure 0). The ninth division under any class is frequently reserved for miscellaneous items which are as yet of too small importance to classify separately; this should not be confused with the first item (0) under each class which is used for general material applied to many or all of the subdivisions under it.

b. Detailed Form Classification

The Dewey classification as well as the extension for radio is mainly by subject or content, regardless of form. For material covering a general field special form subdivision of the subject is found practically useful. For classification as to form the following set of numbers may be used in connection with the number corresponding to any subject covered.

01	Theory; methods; programs.
02	Textbooks; outlines; manuals.
03	Cyclopedias; dictionaries.
04	Essays; addresses; lectures; letters; papers.
05	Periodicals; magazines; reviews; bibliography; publications.
06	Societies; associations; transactions; exhibitions.
07	Education; training; museums.
08	Tables; calculations; charts; maps.
09	History; progress; development; biographical.
001	Statistics.
002	Quantities; cost.
003	Contracts; specifications.
004	Designs; drawings.
005	Executive; administrative; rules.
006	Working; maintenance.
007	Laws; regulations.
008	Patents.
009	Reports of tests; bulletins.

The sequence of figures constituting the form number is simply placed to the right of the sequence of figures constituting the class number. If the class number already ends in one or two zeros, as 500 or 510, these zeros are disregarded in making up the combined number. Thus a periodical on any subject has the subject number followed by 05.

Examples:

R505.	Periodicals on applications of radio.
R510.5	Periodicals on applications of radio to navigation.
R526.105	Periodicals on radio-beacon systems.
R526.100.7	Laws regarding radio beacons.

Thus the classification of any subject may be expanded to meet the needs of an individual file. The complete number gives in a condensed form an indication of what the material is as well as its location in the files.

III. Classification of Radio Subjects

a. DETAILS OF USE

In the classification of radio subjects the main features of the Dewey system as to subject and form classification are retained.

The class (R800) is anomalous. This space in the classification is actually used for non-radio matter. Such material should, however, be given its regular class number according to the Dewey system. If it were arranged in strictly numerical order, some of this material would come before radio and some after radio. By choosing arbitrarily to use the space denoted by R800 for this purpose it is possible to arrange the non-radio material in classified order, but to keep it subordinate to a larger volume of radio material. Accordingly, a number of non-radio items are included where R800 comes in the list under Section IV below but given their number according to the complete classification.

In filing a specific book or paper under a given class or subdivision, its title may be indexed by the classification number plus a small letter, the assignment being according to subject, author, chronological order of accession, or any other consideration depending on the circumstances.

In a card file of references to periodical literature, it is convenient to arrange the cards under each final class or subdivision in chronological order or in alphabetical order by the names of authors.

The needs of individual collections of files vary widely and expansion of the system can be made by any person using the system. The following classification table is given as a classification which in itself meets the needs of small collections or files. Persons interested in a particular subject or subjects will find it advantageous to expand the parts in which they are interested and to use the classification as given for those parts in which they have only a general interest.

In Section IV below there is given a detailed extension of this classification which has been evolved to meet the filing needs of the Radio Section of the Bureau of Standards. In that table there will be found examples of detailed extensions to meet particular circumstances.

In cases where files of an organization are numbered according to an extended system and are made available to another organization using a less extended system, the detailed portion of the classification numbers can be removed. An example of this is the monthly lists of References to Current Radio Literature published by the Bureau of Standards.[3] The reference numbers in the Bureau's own files are according to the table given in Section IV, e.g., an article on radio-beacon systems for aircraft (visual type) is filed under R526.12. This may be filed in a less extended file under R520 (aircraft radio), R526 (radio as navigation aid), or R526.1 (beacon systems for aircraft) depending on how brief a filing system is being used.

b. CLASSIFICATION TABLE

R000	RADIO COMMUNICATION (Material of a general nature for which no specific classification can be used and which relates to the field as a whole).
R100	RADIO PRINCIPLES (Material having to do with underlying theory).
R110	Radio Waves (Transmission phenomena and theory, atmospherics).
R120	Antennas
R130	Electron Tubes
R140	Circuit theory and effects
R150	Generating apparatus
R160	Receiving apparatus
R170	Interference
R190	Other radio principles
R200	RADIO MEASUREMENTS AND STANDARDIZATION (Methods of and apparatus for measurement)
R210	Frequency
R220	Capacity
R230	Inductance
R240	Resistance, current, voltage, etc.
R250	Generating apparatus
R260	Receiving apparatus
R270	Intensity (Field intensity, signal intensity, noise, etc.)
R280	Properties of materials
R290	Other radio measurements
R300	RADIO APPARATUS AND EQUIPMENT (Component parts or apparatus, not complete communication systems)
R310	Antennas
R330	Electron tubes
R350	Generating apparatus
R360	Receiving apparatus
R380	Parts, instruments
R390	Other radio apparatus and equipment
R400	RADIO COMMUNICATION SYSTEMS (Complete communication systems or parts of a system which are considered in relation to the complete system).

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[3] PROC. I.R.E. and Radio Service Bulletin.

R410	Modulated-wave systems
R420	Continuous-wave systems
R440	Remote control (by wire)
R460	Duplex and multiplex systems
R470	Radio-frequency carrier wire systems
R480	Radio relay systems
R490	Other systems
R500	APPLICATIONS OF RADIO
	(Radio as an instrument in other arts and industries)
R510	Marine
R520	Aeronautics
R530	Commercial and special services
	(Commercial communications, press, railroads, mining, etc.)
R540	Private
	(Amateur)
R550	Broadcasting
R560	Military
R570	Remote control by radio
R590	Other applications
R600	RADIO STATIONS
	(Equipment, operation and management)
R610	Equipment
R620	Operation and management
R700	RADIO MANUFACTURING
R710	Factories
R720	Processes
R740	Sales
R800	NON-RADIO SUBJECTS
	(Material of interest but not a part of radio. Give numbers according to the Dewey system.)
R900	MISCELLANEOUS RADIO
	(Material which has no specific place. See R000).

IV. Extensions of Radio Classifications

For larger collections and files a still more detailed extension might be required. The form classification (Sec. II, 2) is very useful for detailed extensions, and may be used under any item in the classification, as occasion requires.

The following extension of the subject classification has been developed for filing material in the Radio Section, Bureau of Standards. Form classifications are not given in the table except under R000, but may be made automatically, anywhere in the classification as occasion requires. Radio reference lists and other material published by the Bureau are classified according to this table.

V. Extended Classification Table as Used in Bureau of Standards Files

The extended classification table as used in the Bureau of Standards files is not reproduced here since it is that given on pages 7 to 21 of the present Circular 138 with only minor changes.

BOOK REVIEW

Radio Engineering Principles, By Henri Lauer and Harry L. Brown. Second Edition. McGraw-Hill Book Company, Inc., 370 Seventh Avenue, New York City. 301 pages, 227 illustrations. Price $3.50.

"Radio Engineering Principles" now appears in a second edition, thoroughly revised in accordance with the latest practice of the radio art. As the title indicates, the authors confine themselves to principles, rather than describing particular instruments and devices, such as broadcast and ship transmitter installations and commercial and broadcast receivers. The theory of radio transmission and reception is unfolded by direct description and simple mathematics rather than by the conventional method of physical analogies.

The first two chapters are concerned with electrical and magnetic phenomena; the third with antenna systems and radiation; the fourth and fifth with damped and undamped wave telegraphy; the balance of the book with vacuum tubes and their application to every phase of transmission and reception. This incomplete summary of the text is given only to show that a proper balance, in the light of the modern importance of the vacuum tube, has been maintained by the authors. This adequate treatment of the vacuum tube and its applications is particularly thorough and scholarly. The quantitative theory of oscillation generation and the theory of push-pull amplification are two examples of especially good mathematical presentation. The authors make generous acknowledgment to many sources, some of them decidedly unfamiliar foreign references.

The book is particularly suited as a text for shorter radio courses, given as training to student electrical engineers, and as a general reference text. It deserves a place in every professional radio engineer's library.

Edgar H. Felix†

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† Writer and Broadcast Consultant, Ridgewood, N. J.

Volume 16, Number 10 *October, 1928*

MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE*

THIS is a monthly list of references prepared by the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for ten cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The articles listed below are not obtainable from the Government. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R 100. RADIO PRINCIPLES

R113 Eckersley, T. L. The polarization and fading of short-wave wireless. *Nature* (London), **121**, p. 707, May 5, 1928. *Experimental Wireless* (London), **5**, p. 397, July, 1928.

(The received rays are shown to be circularly or elliptically polarized, an effect which has not been heretofore observed on long-distance stations just outside the skip distance. Effect of earth's magnetic field in changing the plane of polarization is followed up for very short waves (14-50 m). A unidirectional receiving antenna was used. These phenomena attributed to a double refraction phenomenon producing the circular polarized waves).

R113 Hollingworth, J. The polarization of radio waves. *Proceedings* Royal Society of London, **119A**, pp. 444-464; June, 1928.

(General discussion on polarization of radio waves due to Heaviside layer. A modified method is given for studying the effect of frequencies less than 30 kc).

R120 O'Neill, H. M. Characteristics of certain broadcasting antennas at the South Schenectady development station. PROC. I.R.E., **16**, pp. 872-889, July, 1928.

(Characteristics of various antennas used for broadcasting. Measurements at the station discussed. Effect of signal strength as measured locally for different antenna heights and effect of high steel towers on antennas operated at 380 meters are treated).

R126 Parker, H. Radio grounds for broadcast receivers. *Radio* (San Francisco), **10**, pp. 29-30; August, 1928.

(Data on resistivity of soil for best grounds for broadcast reception).

R130 Prince, D. C. Four-element tube characteristics as affecting efficiency. PROC. I.R.E., **16**, pp. 805-821; June, 1928.

(Study of ratio of grid and plate currents of symmetrical tubes (cylindrical grids and plates) shows that it is quite different from that with ordinary commercial tubes. The difference appears to be due to combination of secondary emission from the plate and unsymmetrical arrangement of the grid wires).

R132 Williams, N. H. The screen-grid tube. PROC. I.R.E., **16**, pp. 840-843; June, 1928.

(Emphasis of paper on the very high amplification (up to eighty times per stage) which can be obtained by use of screen-grid tube. Shows that the current through the tube is approximately independent of the plate voltage and therefore the voltage amplification is given by the product of the mutual conductance and the lead impedance).

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* Original Manuscript Received by the Institute, August 13, 1928.

R134.45 David, M. Super-reaction. (Super-regeneration), *L'Onde Electrique*, **7**, pp. 217-260; June, 1928.

(Discussion of work done to date on super-regeneration).

R 200 RADIO MEASUREMENTS AND STANDARDIZATION

R201 Marrison, W. A. Thermostat design for frequency standards. PROC. I.R.E., **16**, pp. 976-980; July, 1928.

(Special design of thermostatic control for frequency standardization. Thermal system arranged so that the variations reaching the object to be controlled are materially reduced below those existing at responding element. This is accomplished by using a layer of material which attenuates temperature variations between the object to be controlled and the region about the responding element.)

R230 Useful data charts—Inductance, capacity and frequency—shortwave band. *Wireless World and Radio Review*, **23**, pp. 82-83; July 18, 1928.

(Handy chart for calculation of above).

R 300. RADIO APPARATUS AND EQUIPMENT

R330 McLachlan, N. W. The output stage and the Pentode. *Wireless World and Radio Review*, **22**, pp. 30-33; July 11, 1928.

(New electron tube for use as loudspeaker tube. Curves and characteristics of tube are given).

R382 Replogle, D. E. Additional notes on iron core reactances. *QST*, **12**, p. 46; August, 1928.

(Design chart for filter reactors).

R 400. RADIO COMMUNICATION SYSTEMS

R431 Carson, J. R. The reduction of atmospheric disturbances. PROC. I.R.E., **16**, pp. 966-75; July, 1928.

(Analysis of an arrangement which provides for high-frequency selection plus low-frequency balancing after detection for reduction of atmospherics).

R 500. APPLICATION OF RADIO

R520 Hanson, M. P. Aircraft radio installations. PROC. I.R.E., **16**, pp. 921-65; July, 1928.

(Technical aspects of aircraft radio design and installation given. Illustrates trend of development during recent years).

R520.5 Jolliffe, C. B. and Zandonini, E. M. Bibliography on aircraft radio. PROC. I.R.E., **16**, pp. 985-99; July, 1928.

(List of two hundred fifty seven references to domestic and foreign periodicals).

R526.1 Smith-Rose, R. L. Directional wireless and marine navigation rotating loop beacons. *Nature* (London), **121**, p. 745, May 5, 1928. *Experimental Wireless* (London), **5**, p. 402, July, 1928.

(Radio-beacon system of directional wireless transmission developed by the Royal Air Force, Great Britain. This beacon may be of value to marine navigation. At distances over 60 miles over sea night effects change the true bearing somewhat. Rotating loop transmitting system seems to have certain advantages over the rotating loop receiving system).

R526.1 Dellinger, J. H. and Pratt, H. Development of radio aids to air navigation. PROC. I.R.E., **16**, pp. 890-920; July, 1928.

(Technical description of the system developed by the Bureau of Standards on a radio-beacon system and telephone service from ground to aircraft).

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED

September 5, 1928

Transferred to the Fellow grade

New Jersey	**Freehold**, 54 Court Street	Englund, Carl R.
	South Orange, 332 Melrose Place	Hartley, R. V. L.
New York	**New York** City, R. C. A., 66 Broad Street	Beverage, Harold H.
	New York City, 195 Broadway	Blackwell, O. B.
	New York City, 64 Broad Street	Kroger, Fred H.
	New York City, R. C. A., 66 Broad Street	Ranger, Richard H.
	New York City, General Electric Co., 120 Broadway	Young, Owen D.
	Schenectady, General Electric Co.	Baker, W. R. G.

Transferred to the Member grade

Illinois	**Chicago**, 2022 The Engineerr Bldg.	Turner, George S.
Maryland	**Baltimore**, Room 13, Custom House	Sterling, George Edward
Massachusetts	**Aburndale**, 21 Lasell Street	Clapp, James K.
	Cambridge, 10 Wyman Road	Weeks, Paul T.
New Jersey	**Newark**, 200 Mt. Pleasant Avenue	Rypinski, Maurice C.
New York	**Grasmere**, S. I.	Grimes David
	Great Neck, L. I.	Rigney, Douglas
	New York City, 40 Broad Street	Blenheim, William J.
	New York City, R. C. A., 70 Van Cortlandt Park So.	Engel, Francis H.
	New York City, 420 Lexington Avenue	Melhuish, Harold T.
	New York City, R. C. A., 66 Broad Street	Winterbottom, W. A.
Pennsylvania	**East Pittsburgh**, 124 Avenue A	Sutherlin, Lee

Elected to the Member grade

California	**Coronado**, P. O. Box 362	Pierce, Francis E.
	Palo Alto, Federal Teldgraph Co.	Brower, William M.
Dist. of Columbia	**Washington**, Radio Division, Dept. of Commerce.	Downey, W. E.
Illinois	**Chicago**, 2022 The Engineering Bldg.	Sloan, F. V.
New York	**Rochester**, 25 Leighton Avenue	Bickford, Milton F.
	Rochester, 25 Leighton Avenue	Haugh, A. T.
Ohio	**Barberton**, Ohio Insulator Co.	Hillebrand, W. A.
Peru	**Lima, c/o** Marconi's Wireless Telegraph Co., Ltd.	Madgwick, G.

Elected to the Associate grade

Arkansas	**Hot Springs**, 122 Harrell Avenue	Peel, DeLacy J
California	**Los Angeles**, 741½ S. Alvarado	Richards, A. H.
	Los Angeles, 1224 Wall Street	Sorge, Barthold W.
	Palo Alto, 804 Bryant Street	Dow, Clifford J.
	Santa Barbara, c/o Lamb Electric Co.	Cooper, Harold Grant
	Yosemite, c/o Ahwahnee Hotel	Campbell, Wm. N.
Connecticut	**Hartford**, 52 Girard Avenue	Briggs, Perry Orswell
	Hartford, 1711 Park Street	Meserve, G. Donald
	Stamford, Y. M. C. A.	Carlson, Einar C.
	Waterbury, P. O. Box 188	Cohen, Edward
Delaware	**Claymont**	Tatnall, Joseph S.
Dist. of Columbia	**Washington**, 2067 Park Road, N. W.	Gillaland, Ted R.
	Washington, 3320-19th Street, N. W.	Zandonini, Elizabeth M.
Georgia	**Atlanta**, A. T. & T. Co., 193 S. Pryor St.	Bellinger, Andrew L.
	Columbus, 2303 Hamilton Avenue	Lewis, William J.
	Fort Benning, Radio Station	Allen, Otis T.
	Savanah, S. S. *City of Savanah*, Chief Radio Op.	Rushing, Wallace E.
Illinois	**Chicago**, 1648 N. La Salle Street	Cohen, Hans
	Chicago, 1035 Belden Avenue	James, Gerald W.
	Chicago, 711 Buckingham Place	McCullah, Arthur B.
	Collinsville, 307 W. Main Street	Gasparotti, Ralph G.
	Joliet, Box 1037	Crowley, Wm. J., Jr.
	Oak Park, 916 South Lombard Avenue	Roe, Douglas J.
Indiana	**Terre Haute**, 705 So. 20th Street	Brooks, Kenneth Earle
	Valparaiso, 307 Monroe Street	Hershman, J. B.
	Valparaiso, Dodge's Institute	Larrowe, Harvey F.
Iowa	**Sioux City**, 302 Pierce Street	Kerby, Edward J.
Kentucky	**Owensboro**, Ken-Rad Corporation	Wuertz, John L.
	Paducah, P. O. Box 440	Shelton, Raymond W.
Maryland	**Baltimore**, 10 E. Centre Street	Leight, Edward Roy
Massachusetts	**Beverly, c/o** Neptron Corporation	O'Neill, George D.
	Boston, 30, 20 Egleston Street	Wolf, Edwin A.
	Chicopee Falls, Westinghouse Radio Lab.	Brewer, Noble E.
	Dedham, 116 Riverside Drive	Cowan, Edward J.
	East Springfield, 92 Windemere Street	Madsen, Carl J.
	Lynn, 89 New Park Street	Litle, Roy Scott

Michigan	Detroit, 3760 Jefferson Avenue E.	Bailey, Stuart L.
	Detroit, Marine P. O., Steamer *W. H. McGean*	Leonhardt, Charles C.
	Wyandotte, 344 Orchard Street	Steffin, V. J.
Minnesota	Minneapolis, 4512 Harriet Avenue	Sharpless, William M.
	St. Paul, 1795 Lincoln Avenue	Cotton, Richard J.
Missouri	St. Joseph, 635 N. 22nd Street	Bauer, Fritz
	Springfield, 517 E. Walnut Stteet	Hicks, Kenneth F.
New Jersey	Bloomfield, 118 Washington Street	Newell, Guy
	Caldwell, 14 Farrington Street	Wheeler, C. M.
	Camden, 220 Cooper Street	Reeser, Delbert A.
	East Orange, 10 Kensington Place	Clarry, Harold E.
	Fort Hancock	Mallon, William H.
	Glen Ridge, 163 Linden Avenue	Scherer, W. M.
	Hoboken, Stevens Tech	Paulding, Herbert L.
	Jersey City, 276 Summit Avenue	Sullivan, S. Edward
New York	Brooklyn, 2172–72nd Street	Hajim, Jack
	Buffalo, 281 Emslie Street	Roberts, Alfred E.
	Edgemere, L. I., 447 Beach 46 Street	Carduner, William
	Ithaca, Cornell University, Physics Dept	Verman, Lal Chand
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	Toronto, Ontario, 474 Palmerton Blvd	Price, Harold W.
	Winnipeg, 529 Dominion Street	Pound, Harris D.
China	Peking, University of Communications, Electrical Dept	Nieh, C. R.
England	Bath, Kennington Road, Warwick Villa	Young, Alfred W.
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	Caversham, Reading, 4 Matlock Road	Hill, Ernest N.
	Frinton-on-Sea, Shirley, Old Road	Harries, John Henry O.
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	London, N. W. 3, 85 King Henry's Road	Ferdinando, Christie
	Upminster, Essex, "Hilly's," Cedar Gardens	Jones, John E. Rhys.
India	Hyderabad Sind, Gidu Road, Porch House	Samtani, R. C.
New Zealand	Wellington, 248 The Terrace	Taylor, James O.

Elected to the Junior grade

California	Los Angeles, 219 South Alexandria Avenue	Dawson, W. Robert
Connecticut	Mystic, 21 W. Mystic Avenue	Wheeler, Lester B.
Illinois	Chicago, 3128 Warren Avenue	Demikis, Anton, Jr.
	Chicago, 2735 N. Tripp Avenue	Wegner, Charles W.
	Pittsfield, Box 514	Lewis, Rupert S.
Indiana	Valparaiso, 156 S. Franklin Avenue	Barber, D. G.
Iowa	Clarinda, 514 Main Street	Huntsinger, P. R.
Massachusetts	Cambridge, 11 Carlisle Street	Sullivan, Edwin T.
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	Brooklyn, 1819 Newkirk Avenue	Rippere, R. Oliver
	New York City, 1297 Lexington Avenue	Bruggeman, John T.
	New York City, U. S. S. *Gresham*, Barge Office Rm. 10	Mandell, Benjamin F.
Oregon	Falls City	Paul, Byron R.
Pennsylvania	Ambridge, 303 Maplewood Avenue	Svegel, Peter J.
	Harrisburg, 1712 Forster Street	Chambers, William B.
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APPLICATIONS FOR MEMBERSHIP

Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Committee on Admissions. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before November 3, 1928. These applications will be considered by the Board of Direction at its November 7th meeting.

For Transfer to the Fellow grade

New York	New York City, 282 West End Avenue	Dreher, Carl

For Transfer to the Member grade

California	Hollywood, KMTR Radio Corp., 1025 North Highland Avenue	Van Why, Forbes William
Illinois	Chicago, University of Chicago, Ryerson Lab.	Hoag, J. Barton
New Jersey	Madison, 14 Fairwoods Road	Manning, Charles T.
England	London, W. 12, Emlyn Road	Hardy, Harold

For Election to the Member grade

Dist. of Columbia	Washington, 4302 Brandywine St., N. W.	Davis, Thomas McL.
Massachusetts	W. Somerville, 53 Francesca Avenue	Kolster, Charles C.

For Election to the Associate grade

Alabama	Phenix City	Moore, C. M.
California	Berkeley, 2112 Addison	Howard, L. W.
	Burlingame, 908 Morrell Avenue	Michalic, Charles Tom
	Hollywood, 1301 Martel Avenue	Kratz, C. F.
	Marshall, c/o Radio Corporation of America	Lawrence, Lloyd H.
	Oakland, 643 Poirier St.	O'Brien, Daniel L., Jr.
	Orange, P. O. Box 145	Charbonneau, L. H.
	San Francisco, 140 New Montgomery, Room 1024	Stewart, Ronald B.
	San Francisco, KFWI, 1182 Market St., Room 203	Weir, F. Torres
	San Pedro, U. S. S. *Procyon*	Granum, Alfred M.
	Santa Ana, 1334 S. Parton St.	Sleeper, J. Lloyd
Georgia	Atlanta, 684 Durant Place	Gue, E. M.
	Atlanta, c/o Radio Station WSB	Shropshire, A. W.
	Columbus, 731 Broad St.	Cowart, Frank P.
	Columbus, South Georgia Power Co.	Davis, James C.
Illinois	Chicago, 3814 Madison St.	Haire, A. F.
	Chicago, 1119 So. Wood St.	Korbel, George William
	Chicago, c/o R. C. A., 100 W. Monroe St.	McCune, R. H.
	Chicago, 4137 Jackson Blvd.	Strandjord, Millard F.
	Chicago, 29 No. Morgan Street	Trier, Stanley S.
	Rantoul, Chanute Field, Communications Section	Fiechtner, Theodore G.
	Rockford, 1506 School St.	Johnson, Art A.
Kansas	Atchison, 202 N. 2nd St.	Dimond, Benjamin Dan
Louisiana	New Orleans, 528 N. Hennessy St.	Salzer, Herman M.
Maine	Belfast, c/o R. C. A.	Morris, Hedley B.
Massachusetts	Braintree, 149 Hollis Ave.	Robinson, Philip F.
	Cambridge, Harvard University, Pierce Hall, Room 204B	Samoiloff, Leon
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	Chathamport, Radio Hotel	Lee, Walter O.
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	Medford, 16 Higgins Avenue	Yorks, William Ray
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Australia	Melbourne, East Camberwell, 6 Beech St.	Fitts, Rupert Alfred
Canada	Victoria, B. C., 2084 Newton St.	Hawkins, Ernest
	Ottawa, 21 Florence St.	Donaldson, Bruce W.
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For Election to the Junior grade

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Nebraska	Clay Center	Hertel, Roger H.
New York	Brooklyn, 6 Bay 23rd St.	Liebner, Barney
	Buffalo, 78 East St.	Ferger, Herbert
	Jamestown, 200 Hallock St.	Ellis, James G., Jr.
	Lancaster, 69 Aurora St.	Pictor, Robert Ezra
Pennsylvania	Altoona, 602 E. Grant Ave.	Youngkin, E. E.
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	Scranton, 1618 Monsey Ave.	Batzel, Charles R.
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England	Manchester, Trafford Park, Metro-Vickers	Thomas, Guy Henry
	Wiltshire, Swindon, 7, Pinehurst Road	Humphreys, Leonard W.

PROCEEDINGS OF
The Institute of Radio Engineers
Volume 16 November, 1928 Number 11

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GENERAL INFORMATION

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ARTHUR BATCHELLER
Member of the Board of Direction of the Institute, 1928

Arthur Batcheller

MEMBER OF THE BOARD OF DIRECTION OF THE INSTITUTE, 1928

Arthur Batcheller was born at Wellesley, Massachusetts, in 1888. He studied electrical engineering for three years at the Y.M.C.A. Polytechnic School at Boston, Mass., and was instructor for four years in radio and electricity at the evening sessions of the Boston School of Telegraphy. Mr. Batcheller was assistant to the electrical engineer in charge of installation and maintenance of railroad electric signal system for the Boston, Revere Beach, and Lynn Railroad for four years. From 1915 to 1916 he was employed as chief electrician for the Eastern Steamship Corporation at Boston, and from 1917 to 1919 he was U. S. Radio Inspector in charge of the New England District. In 1919 he became associate partner and founder of the Massachusetts Radio Telegraph School. Mr. Batcheller accepted an appointment as Radio Inspector in charge of the Second Radio District in 1920 with headquarters in New York. He has served as Supervisor of Radio since that date.

In 1925 he was commissioned a Lieutenant Commander in the United States Naval Service.

Mr. Batcheller is a member of the Committee on Conference to study safety of life at sea to be held in London in 1929. He was elected to Associate membership in the Institute in 1914 and was transferred to the Member grade in 1920.

Mr. Batcheller was appointed a member of the Board of Direction by the Board for a one year term in January of 1928.

CONTRIBUTORS TO THIS ISSUE

Bashenoff, Valerian I.: Born January 9, 1890 at Kostroma, U.S.S.R. E. E. degree, Polytechnical High School, 1918; in charge of radio engineering, 1918–1920. Engineer in charge at State Radio Works, Moscow, 1920–1921. Docent in radio engineering at Moscow Technical High School, 1921; associate professor of radio engineering at the Moscow Transport Engineers Institute, 1924–1928. Charter Member (1918) and Secretary (1919–) of the Russian Society of Radio Engineers. Secretary of the All-Russian Radio Association, 1921–1925. Since 1921 chief engineer of the radio department of the State Electrical Research Institute. Since 1927 technical editor of radio engineering section of the *Technical Encyclopedia.* Member of the American Association for the Advancement of Science; Member of the Institute of Radio Engineers, 1927.

Colwell, Robert C.: Born October 14, 1884 at Fredericton, N. B., Canada. A. B. degree, Harvard University; A. M. degree, University of New Brunswick; Ph. D. degree, Princeton University; professor of physics, Geneva College, 1913–1923; assistant director of radio laboratory, West Virginia University, 1918–1919; professor of physics, West Virginia University, 1924– . Member American Physical Society, Franklin Institute, American Mathematical Society. Associate member, Institute of Radio Engineers, 1921.

Dellinger, J. H.: (See Proceedings for May, 1928.)

Dickey, Edward T.: Born November 16, 1896 at Oxford, Penna. B. S. degree, College of the City of New York, 1918. Amateur experimenter, 1908–1918. Marconi Wireless Telegraph Company of America and Radio Corporation of America, 1918 to date. Chairman of the Subcommittee on Receiving Sets of the Standardization Committee, I.R.E., and member of the Meetings and Papers Committee, 1928. Fellow in the Radio Club of America; Junior member, Institute of Radio Engineers, 1915; Associate, 1917; Member, 1923.

Engel, Francis H.: Born July 19, 1899 at Washington, D. C. Educated at George Washington University, Washington, D. C. With National Electrical Supply Company, 1917–1918; U. S. Naval Radio Service, 1918–1919; radio laboratory, U. S. Bureau of Standards, 1919–1923; Radio Corporation of America, 1923 to date. Associate member, Institute of Radio Engineers, 1925; Member, 1928.

Harrison, J. R.: Born September 26, 1903 at Boston, Massachusetts. B. S. degree, Tufts College, 1925; M. A. degree, Wesleyan University, 1927. Electrician, Boston and Maine Railroad, 1921–1925. Assistant in physics, Wesleyan University, 1925–1927. Charles A. Coffin Fellow (Charles A. Coffin Foundation established by the General Electric Company) at Wesleyan University, 1927– . Member, Optical Society of America and American Physical Society; Associate member, Institute of Radio Engineers, 1926.

Schelleng, J. C.: Born November 11, 1892 at Freeport, Ill. A. B. degree, Cornell University, 1915. Instructor in physics at Cornell, 1915–1918. Since 1918 with the Western Electric Company and Bell Telephone Laboratories working on radio transmitters and transmission. For past few years in charge of studies on high power transmitters, transmitting antennas, and radio transmission. Associate member, Institute of Radio Engineers, 1923; Member, 1925; Fellow, 1928.

Terry, Earle M.: Born January 16, 1879 at Battle Creek, Michigan. A. B. degree, University of Michigan, 1902; M. A. degree, University of Wisconsin, 1904; Ph. D. degree, University of Wisconsin, 1910. Instructor, University of Wisconsin, 1904–1910; assistant professor, 1910–1917; associate professor, 1917–1928; professor, 1928, in charge of teaching engineering physics. Member of research staff, U. S. Naval Experimental Station during World War. In charge radio station WHA, University of Wisconsin. Previous research work in field of magnetism. President, local chapter Sigma Xi. Fellow, American Physical Society; Associate member of the Institute of Radio Engineers, 1919; Member, 1928.

Turner, H. M.: (See PROCEEDINGS for June, 1928.)

Van Dyck, A. F.: Born May 20, 1891 at Stuyvesant Falls, New York. Ph. B. degree, Sheffield Scientific School, Yale University, 1911. Amateur experimenter and commercial operator at sea, 1907–1910. With National Electric Signalling Company, Brant Rock, Mass., 1911–1912; Westinghouse Electric and Manufacturing Company, Research Dept., 1912–1914; instructor in electrical engineering, Carnegie Institute of Technology, 1914–1917; expert radio aide, U. S. Navy, 1917–1919; Marconi Company, General Electric Company, Radio Corporation of America, 1919 to date. Charter Associate member of the Institute of Radio Engineers; Member, 1918; Fellow, 1925.

INSTITUTE NOTES AND RELATED ACTIVITIES

OCTOBER MEETING OF THE BOARD OF DIRECTION

A meeting of the Board of Direction of the Institute was held in the Institute offices at 4 P.M. on October 3rd, 1928. The following were present: Alfred N. Goldsmith, President; L. E. Whittemore, Vice-President; Melville Eastham, Treasurer; Arthur Batcheller, W. G. Cady, J. H. Dellinger, R. A. Heising, R. H. Manson, R. H. Marriott, and John M. Clayton, Secretary.

Upon recommendation of the Committee on Admissions, the following were transferred or elected to the higher grades of membership in the Institute: transferred to the grade of Fellow: Arthur E. Kennelly and C. W. Rice; elected to the grade of Fellow: Captain S. C. Hooper and Lynde P. Wheeler; transferred to the grade of Member: Kenneth N. Cumming and Alfred H. Hotopp; elected to the grade of Member: K. R. Smith and H. J. Walls.

Sixty Associate members and six Junior members were elected.

1928 MORRIS LIEBMANN MEMORIAL PRIZE

Upon the recommendation of the Committee on Awards, the Board of Direction decided that the 1928 Morris Liebmann Memorial Prize is to be awarded to Professor Walter G. Cady, of Wesleyan University, for his fundamental investigations in piezoelectric phenomena and their application to radio technique.

The vote of the Committee on Awards and the Board was unanimous save for Dr. Cady's dissent in each case.

It was decided that the award is to be made at the November 7th meeting in New York City.

The Morris Liebmann Memorial Prize has been awarded in past years as follows:

Leonard F. Fuller	1919
R. A. Weagant	1920
R. A. Heising	1921
C. S. Franklin	1922
H. H. Beverage	1923
John R. Carson	1924
Frank Conrad	1925
Ralph Bown	1926
A. Hoyt Taylor	1927

BOARD OF EDITORS

To assist in the greatly increased work involved in the editing of manuscripts received for publication in the PROCEEDINGS, a Board of Editors of the PROCEEDINGS has been appointed with membership as follows: Alfred N. Goldsmith, Chairman; Stuart

Ballantine, Ralph Batcher, W. G. Cady, Carl Dreher, and G. W. Pickard.

The work of the Committee on Meetings and Papers has not been altered through the appointment of the Board of Editors. The Committee will continue to function as an Institute body which secures papers for publication and passes upon their suitability from a technical standpoint for publication in the PROCEEDINGS.

1929 YEAR BOOK

The membership list in the 1929 Year Book will be dated as of December 15th, 1928. No changes of address can be included in the 1929 Year Book unless such changes are received by the Institute not later than December 15th.

Members who have failed to return the address slips which accompanied the ballots for 1929 officers and members of the Board, mailed early in October, are urged to do so immediately.

It is expected that the 1929 Year Book will contain the adopted report of the Committee on Standardization in addition to the material published in former Year Books.

CHANGES OF ADDRESS

In future years it is planned that no "change of address" form will be sent to the entire membership each year. The addresses for the membership list of each Year Book will be kept up to date continuously.

Members who change their business connections or business title, even though there is no change in mailing address, are requested to advise the Institute office when such changes are made so that the Year Book catalog may correctly list both their mailing and business addresses.

FORTHCOMING PAPERS

The following is a list of papers on hand for probable publication in the PROCEEDINGS. It is expected that these papers will be published in early forthcoming issues:

Note on the Effect of Reflection by the Microphone in Sound Measurements, by Stuart Ballantine.

The Design of Transformers for Audio-Frequency Amplifiers with Preassigned Characteristics, by Glenn Koehler.

A Fine Adjustment Frequency Control for R. F. Oscillators, by G. F. Lampkin.

A Bridge Circuit for Measuring the Inductance of Coils While Passing Direct Current, by V. D. Landon.

Detection Characteristics of Three-Element Vacuum Tubes, by F. E. Terman.

A Note on the Directional Observations of Grinders in Japan, by E. Yokoyama and T. Nakai.

STANDARD FREQUENCY TRANSMISSIONS BY THE BUREAU OF STANDARDS

The Bureau of Standards' Standard Frequency Transmissions for the months of November, 1928 to March, 1929 inclusive are given as follows:

Standard Frequency Schedule of Frequencies in Kilocycles

Eastern Stand. Time	Nov. 20	Dec. 20	Jan. 21	Feb. 20	March 20
10:00–10:08 P.M.	1500	4000	125	550	1500
10:12–10:20 P.M.	1700	4200	150	600	1700
10:24–10:32 P.M.	2250	4400	200	650	2250
10:36–10:44 P.M.	2750	4700	250	800	2750
10:48–10:56 P.M.	2850	5000	300	1000	2850
11:00–11:08 P.M.	3200	5500	375	1200	3200
11:12–11:20 P.M.	3500	5700	450	1400	3500
11:24–11:32 P.M.	4000	6000	550	1500	4000

Members are referred to page 1300 of the October, 1928 issue for more detailed information regarding these signals from the Bureau's Station WWV at Washington, D. C.

Information on how to receive and utilize the signals is given in Bureau of Standards' letter circular No. 171, which may be obtained by applying to the Bureau of Standards, Washington, D. C.

Institute Meetings

ATLANTA SECTION

A meeting of the Atlanta Section was held in the Chamber of Commerce Building, Atlanta, on September 13th at 8 P.M. Walter Van Nostrand, chairman of the Section, presided. The meeting was addressed by Irving Wolff, of the Radio Corporation of America, on the subject of "Sound Measurements and Loud Speaker Characteristics," which included a brief description of the method used in measuring loud speaker response, comparing the Rayleigh Disk and Condenser Microphone as sound detectors. A number of loud speaker sound pressure response curves were shown and interpreted in terms of pleasantness of reproduction as determined by low and high frequency cut-off, smoothness of response and tone balance. The speaker also discussed tube overloading and the effect of loud speaker response on its apparent accentuation and diminution. The effect of room absorption characteristics, room resonances, position of loud speaker in the room, and position of listener with respect to the loud speaker were discussed. Numerous lantern slides showing diagrams and loud speaker response curves were shown.

Twenty-five members of the Section attended the meeting.

CANADIAN SECTION

The Canadian Section held a meeting on September 14th in the auditorium of T. Eaton Company. A. M. Patience presided. J. L. McCoy, of the Westinghouse Electric and Manufacturing Company of East Pittsburgh, Pa., presented a paper on "The Televox."

Messrs. Patience, Pipe, Soucy, and others participated in the discussion which followed.

One hundred and fifty members of the Section and their guests attended the meeting.

CLEVELAND SECTION

The October meeting of the Cleveland Section was held on the 10th of the month in the Case School of Applied Science. John R. Martin, chairman of the Section, presided.

The paper of the evening was presented by R. H. Ranger, of the Radio Corporation of America, the subject being "Recent Developments in Photo Radio."

The paper included a demonstration of telephoto and radio photograph transmission as well as numerous lantern slides.

Three hundred and thirty-five members of the Section and guests attended the meeting.

The next meeting of the Cleveland Section will be held on Friday, November 23rd, in the Case School of Applied Science. Dayton C. Miller will present a paper, "Analysis of Musical Sounds," and John R. Martin will present a paper on "Experimental Analysis of Loud Speaker Input and Output."

CONNECTICUT VALLEY SECTION

On September 13th in the Hotel Garde at Hartford, Conn., the meeting of the Connecticut Valley Section was addressed by Edward T. Dickey and Francis H. Engel. K. S. Van Dyke, vice-chairman of the Section, presided. Mr. Dickey's paper was "Quantitative Measurements Used in Testing Broadcast Receiving Sets," and Mr. Engel's, "Vacuum-Tube Production Tests."

Both papers are published elsewhere in this issue of the PROCEEDINGS. Following the presentation of the papers a general discussion took place. Twenty-five members of the Section attended the meeting.

LOS ANGELES SECTION

A meeting of the Los Angeles Section was held on September 17th in the Elite Cafe, South Flower Street, Los Angeles. D. C. Wallace, chairman of the Section, presided.

L. Elden Smith read a paper on "The Problems and Methods of Television Transmission," which was followed by a paper by R. P. Parrish on "The Problems and Methods of Television Reception."

Preceding the meeting an informal dinner was held.

Forty-seven members and guests attended the meeting.

NEW ORLEANS SECTION

The first formal meeting of the newly organized New Orleans Section was held on September 15th at New Orleans.

Dr. Wolff presented a paper, "Sound Measurements and Loud Speaker Characteristics." Messrs. Andres, DuTreil, Lehde, and Gardberg took part in the discussion which followed the paper. Officers of the Section were elected as follows: Pendleton E. Lehde, chairman; L. J. N. DuTreil, vice-chairman, and Anton A. Schiele, secretary-treasurer.

On October 8th a meeting of the New Orleans Section was held in New Orleans. Chairman Pendleton E. Lehde presided.

Two papers were presented; the first by Mr. Lehde was entitled "The Underlying Principles of the Vitaphone and Movietone." The second paper was by Anton A. Schiele on "The Basic Principles of Radio Television." The following members participated in the discussion which followed: J. M. Roberts, J. M. Shaw, C. Schneider, D. S. Elliott.

NEW YORK MEETING

A meeting was held in New York City on October 3rd in the Engineering Societies Building, 33 West 39th Street. Alfred N. Goldsmith, President of the Institute, presided.

A paper was read by J. R. Harrison, of Wesleyan University, on "Piezo-Electric Oscillator Circuits with Four-Electrode Tubes." Messrs. Goldsmith, Harrison, Cady, and Hund participated in the discussion which followed.

The paper is printed elsewhere in this issue of the PROCEEDINGS.

Three hundred and twenty-five members of the Institute attended this meeting.

On December 5th a New York meeting will be held in the Engineering Societies Building, 33 West 39th Street. H. H. Beverage and H. O. Petersen will present a paper "Recent Short Wave Developments."

PHILADELPHIA SECTION

The Philadelphia Section held a meeting in the Franklin Institute on October 10th. J. C. Van Horn, chairman of the Section, presided. Austin Bailey, of the American Telephone and Telegraph Company, read a paper, "The Receiving System for Long Wave Transatlantic Radio Telephony."

A general discussion on the part of the thirty members of the Section present followed the presentation of the paper.

This paper is published in this issue of the PROCEEDINGS.

SEATTLE SECTION

On August 28th a noonday luncheon for members of the Seattle Section was held. F. M. Ryan, of the Bell Telephone Laboratories, presented a paper, "Recent Developments in Radio Broadcasting." Thirty members of the Section and guests attended the luncheon and participated in the discussion of Mr. Ryan's remarks.

On September 19th a luncheon was held with W. K. Bert as the guest. Mr. Bert outlined the recent broadcasting frequency assignments. His remarks were supplemented by a general discussion on this and allied subjects.

A meeting of the Seattle Section was held in the Navy Yard at Puget Sound, Washington, on September 29th. Lieutenant Haas, radio officer of the USS *Lexington*, presented a paper on "Radio Communication Applied to Naval Aviation." The address given outlined the communication organization of the Navy and covered in detail the various types of equipment and systems used. A description of the method of operation and results obtained from the fathometer, radio compass, submarine telegraph, and range finder in addition to some of the radio telegraph systems was given.

Messrs. Austin, Deardorff, Renfro, and others discussed the paper. Preceding the presentation of the paper the fifty members of the Section present inspected the USS *Lexington* under the direction of Lieutenant Haas.

Oliver C. Smith, secretary-treasurer of the Seattle Section, has been transferred to Spokane, Washington. Abner R. Wilson, of Seattle, has been appointed secretary-treasurer to succeed Mr. Smith.

WASHINGTON SECTION

On October 11th in the Continental Hotel, North Capitol Street, Washington, D. C., a meeting of the Washington Section was addressed by Austin Bailey, of the American Telephone and Telegraph Company, on "The Receiving System for Long Wave Transatlantic Radio Telephony." F. P. Guthrie, chairman of the Section, presided.

Preceding the meeting forty-nine members and guests attended an informal dinner. The attendance at the meeting was seventy-two.

In the discussion which followed the paper the following took part: Messrs. A. H. Taylor, A. Hund, H. G. Dorsey, G. D. Robinson, G. Howard, F. P. Guthrie, and E. B. Dallin.

On November 8th a meeting of the Washington Section will be held in the Continental Hotel at which time Warren B. Burgess, of the Naval Research Laboratory, will present a talk on "The Radio Compass in Theory and Practice."

Committee Work

COMMITTEE ON CONSTITUTION AND LAWS

The Committee on Constitution and Laws held a meeting in the offices of the Institute on October 2nd. R. H. Marriott, chairman; G. W. Pickard, and H. E. Hallborg were present.

The Committee is continuing the work in connection with the revision of the Institute's Constitution and By-Laws.

COMMITTEE ON STANDARDIZATION

A meeting of the Committee on Standardization was held in the offices of the Institute on October 2nd, 1928. The following members were present: L. E. Whittemore, Chairman; M. C. Batsel, W.G. Cady, J.H. Dellinger, E. T. Dickey, W. E. Holland, C. B. Jolliffe, R. H. Manson, E. L. Nelson, A. F. Rose, W. J. Ruble, and H. M. Turner.

The Committee began the consideration of the material transmitted to the members with the letter calling this meeting. Action was taken on all definitions contained in sections 1 to 5, inclusive, of the preliminary draft report and on the section on electron tube nomenclature.

Copies of material submitted by the Subcommittee on Bibliography and the Subcommittee on the Use of the Transmission Unit in Radio were distributed. The general plan being followed by these two subcommittees was approved.

It was agreed that the Subcommittee on the Use of the Transmission Unit in Radio should become known as the Subcommittee on Radio Transmission, and that there should be added to the matters which it has under consideration such questions as the following:

Definitions of terms appearing in the section on "Antennas."

Measurement of radiation and specification of service area of broadcasting stations.

Constancy of frequency of transmitting stations.

Specification and measurement of harmonic radiation from transmitting stations.

Specification and measurement of transmission-frequency characteristic-fidelity of transmission.

Specification and measurement of modulation.

It was agreed that further meetings of the committee should be held at approximately two-week intervals, in order to complete the consideration of material which is ready for action.

COMMITTEE ON ADMISSIONS

At the meeting of the Committee on Admissions, held in the offices of the Institute on October 10th, R. A. Heising, Chairman; F. H. Kroger, and E. R. Shute were present.

The Committee acted upon ten applications for transfer or election to the higher grades of membership in the Institute.

COMMITTEE ON INSTITUTE-AWARDS

The Board of Direction of the Institute has appointed a standing Committee on Awards with membership as follows: J. V. L. Hogan, Chairman; L. W. Austin, Ralph Bown, W. G. Cady, and A. Hoyt Taylor.

This Committee is to make appropriate recommendation to the Board of Direction for awards of the Institute Medal of Honor and the Morris Liebmann Memorial Prize.

Personal Mention

R. M. Owen, of the Radio Corporation of America, has been transferred to Dallas, Texas.

Elmer L. Brown has recently become service engineer of the California-Victor Distributing Company, of San Francisco.

A. Norwood Fenton has recently become associated with the Sound Department of Metro-Goldwyn-Mayer Studio at Culver City, Cal.

John Q. Gaubert has left the Ward Leonard Electric Company to be in charge of production of the International Resistance Company, of Philadelphia.

J. Warren Horton, until recently associated with the Bell Telephone Laboratories as research engineer, is now connected with the General Radio Company, of Cambridge, Mass.

J. Warren Wright, recently associated with the Naval Research Laboratory, has been transferred to the Design Section, Radio Division, Bureau of Engineering, Navy Department, Washington.

Edwin W. Lovejoy, for a number of years in the service of the Department of Commerce at San Francisco as radio inspector, has been promoted to the position of U. S. Supervisor of Radio at Seattle, Washington.

Oliver C. Smith, of the Pacific Telephone and Telegraph Company, has been transferred to Spokane, Washington as District Transmission Engineer. Mr. Smith has served for some time as Secretary of the San Francisco Section.

Joseph H. Phillips, Jr. has gone to London, England to become technical advisor and consulting engineer to British Phototone, Ltd. Mr. Phillips was formerly associated with the Fox Case Corporation of New York City.

Volume 16, Number 11 *November, 1928*

PIEZO-ELECTRIC OSCILLATOR CIRCUITS WITH FOUR-ELECTRODE TUBES*

BY

J. R. HARRISON

(Charles A. Coffin Fellow, Wesleyan University, Middletown, Connecticut)

Summary—*Two new piezo-oscillator circuits using the screen-grid tube are described. One circuit uses feedback through the crystal, which has two pairs of electrodes with connections to the anode, control grid, and filament of the tube. In the other circuit the two electrodes of the crystal are connected between the anode and control grid. The new circuits are unusually stable and at low frequencies give greater power output than the three-electrode tube circuits. They are particularly useful at those low frequencies for which flexural vibrations are employed. When oscillating at flexural vibration frequencies crystals show a tendency to creep lengthwise until a position is reached which gives maximum power output. If the crystal is displaced from this position it will return to it again.*

THE first piezo-electric oscillator circuit was that described by Cady.[1] In this circuit the quartz plate has a mounting consisting of two pairs of metallic electrodes, *AC* and *BD* (Fig. 1). One pair of electrodes *BD* are connected to the input and the other pair *AC* to the output of the vacuum-tube circuit. The electrodes *C* and *D* are both connected to the filament of the vacuum tube, and they are sometimes replaced by a single large electrode covering the whole side of the crystal. The oscillations are sustained in this circuit by the energy feedback through the vibrating crystal. This circuit has never had wide application because it cannot be used for crystal oscillations at

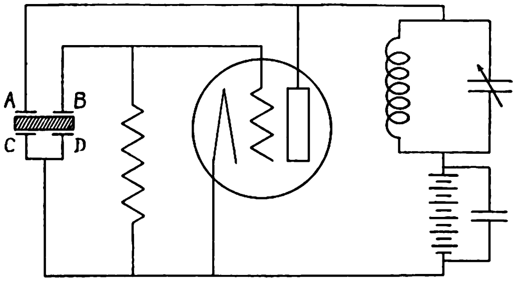

Fig. 1—Well-Known Type of Piezo-Oscillator Circuit Using Feedback through the Crystal.

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* Original Manuscript Received by the Institute, August 1, 1928. Presented at New York Institute Meeting, October 3, 1928.

[1] W. G. Cady, PROC. I. R. E., 10, 111; April, 1922.

the high-frequency modes, that is, the vibrations at the frequency determined by the thickness of the plate.

The writer has found that this Cady type of crystal oscillator may be directly applied to the screen-grid type of vacuum tube.[2] This new circuit which is represented in Fig. 2 possesses several marked advantages over the Pierce type of crystal oscillator now commonly used. Here, as in the Cady circuit, the crystal mounting consists of four electrodes *A*, *B*, *C*, and *D*. The electrodes *C* and *D* are connected to the filament of the tube, and the electrodes *A* and *B* are connected to the plate and control grid respectively. The screen grid *S* is given a positive potential bias equal to about one-third of the plate potential. The plate-potential supply is conveniently used for this purpose. The plate elements *C*, *L* as usual are tuned approximately to the crystal frequency. The chief advantages of this circuit over the Pierce type of oscillator are more constant frequency, greater stability, and, at the lower radio frequencies, greater power output. This circuit may also be used to advantage at the high-frequency modes of the crystal, a performance which as noted previously was not possible with the original Cady oscillator circuit.

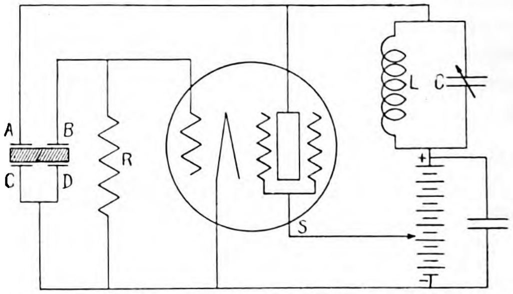

Fig. 2—Piezo-Oscillator with Screen-Grid Tube Similar in Operation to the Circuit of Fig. 1.

POWER TESTS

Power tests have been made of the piezo-electric oscillator as illustrated in Fig. 2. A 7½-watt shielded-grid tube was used with a plate potential of 425 volts and a filament potential of 7½ volts. This tube had a rated maximum plate potential of 500 volts, but only 425 volts were used in order to allow comparison under identical conditions with the three-electrode type of 7½-watt tube, the maximum plate potential of which is of this latter value.

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[2] For the theory and application of the screen-grid, four-electrode tube see the following:
A. W. Hull, and N. H. Williams, *Phys. Rev.*, **27**, 432; 1926.
A. W. Hull, *Phys. Rev.*, **27**, 439; 1926.
J. C. Warner, PROC. I. R. E., **16**, 424; April, 1928.

It has been found that at any given frequency the radio-frequency power output is a maximum at a particular shield-grid potential. Hence in each individual case the shield-grid potential should be adjusted to the optimal value. If this potential is made either too large or too small the power output will be diminished. This shield-grid potential is not very critical, but it is always preferable to make it smaller than the optimal value rather than larger, since the circuit is then more stable.

Tests were made using a quartz plate 30×10×1 millimeters at the transverse effect frequency 92 kc (3300 meters). The power output was 0.55 watts. This same crystal in a Pierce circuit with a UX-210 tube gives a power output of but 0.20 watts. The power output at the lower radio frequencies is always comparatively small, but the tests indicate the relative merits of the two circuits. The optimal shield-grid potential in the above test was 200 volts.

Using the same crystal at the flexural vibration frequency[3] of 50 kc (6000 meters) with the shield-grid tube, the power output was 1.1 watts. The Pierce circuit under identical conditions but with the UX-210 tube gives 0.50 watts power output. The optimal shield-grid potential at this frequency was 225 volts.

In these tests the control grid of the shield-grid tube was given a negative bias by connecting to the negative filament lead through a three-megohm resistor R (Fig. 2). The power output could be considerably increased by substituting a choke coil and bias battery in series for the resistor R. Since this practice of using a choke coil has been found to cause undesirable fluctuations in frequency with three-electrode tube circuits[4] it was not adopted in the final tests cited here. At higher frequencies the power output of the four-electrode tube quartz-oscillator circuit is about the same as that from the Pierce oscillator at the same frequency. This would be expected since the voltage amplification of the four-element tube diminishes as the frequency increases. Hull[5] has found the voltage amplification of certain tubes of this type to be 200 at 50 kc and only 7 at 10,000 kc.

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[3] J. R. Harrison, Proc. I. R. E., 15, 1040; December, 1927.

[4] R. C. Hitchcock, *Electric Journal*, 24, 430; 1927.

STABILITY

It is well-known that in the case of any piezo-electric oscillator any slight damping of the crystal such as the application of pressure from the finger will cause an abrupt termination of oscillations. Furthermore oscillations will not be resumed on removing the cause of the damping until the circuit has been retuned. The screen-grid crystal-oscillator circuit of Fig. 2 is much more stable, for when the circuit has once been tuned for maximum power output, oscillations may be obtained at any time without retuning the circuit elements. For any given frequency a fixed condenser of the proper value could then be conveniently used in this circuit. Oscillations will of course terminate in this circuit when the crystal is sufficiently damped, but they will immediately be resumed when the cause of the damping is removed. The advantages of such characteristics do not need further elaboration here. In order to obtain this high degree of stability fairly good crystals must be used. In many cases crystals which would not oscillate in the three-electrode tube circuit will function in this circuit, but not with the same degree of stability.

In connection with these stability tests an interesting translational movement was noted at the flexural vibration frequency. The rotational movements of piezo-electric quartz plates in high-frequency fields have been described by Meissner[6] and certain translational movements were observed by Hirschhorn.[7] The phenomena here described were observed with a piezo-electric oscillator, whereas Hirschhorn's experiments were with a piezo-electric resonator. If a shield-grid crystal oscillator (Fig. 2) is tuned to maximum power output the tendency of the crystal to emerge from its mounting is observed. But if while oscillating the crystal is partly withdrawn from the mounting and then released it will immediately return to the original position. Sometimes when the crystal is being withdrawn from the mounting the damping is sufficient to kill oscillations, but if the plate has not been withdrawn too far it will again start oscillating and return to the former position.

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[5] A. W. Hull, *Phys. Rev.*, 27, 439; 1926.

[6] A. Meissner, *Zeit. f. Tech. Physik*, 12, 585: 1926. *Zeit. f. Hochfreq.*, 29, 20; 1927. PROC. I. R. E., 16, 281; April, 1927.

[7] S. J. Hirschhorn, *Zeit. f. Physik*, 44, 223; 1927.

If the circuit is not quite tuned to maximum power output the crystal has been found to have a tendency to slide out of the mounting. As the crystal recedes from the mounting the power output increases and the motion ceases when the power output becomes a maximum. As before the limit of emergence, until the circuit stops oscillating, seems to be about one-half the total surface of the plate. It is a curious fact, however, that the maximum power output is the same with one-half of the crystal emerging as it was in the previous case when the circuit was initially tuned to maximum power with the crystal wholly between the electrodes. If now the crystal is forced back between the electrodes the power output will diminish and the crystal will slide out again to the position of maximum power when the restoring force is removed. The crystal is indifferent to the direction of emergence from the mounting; slight inclinations from the horizontal of the crystal mounting seem to be the sole determinant of this direction. In any case, for a given circuit condition the

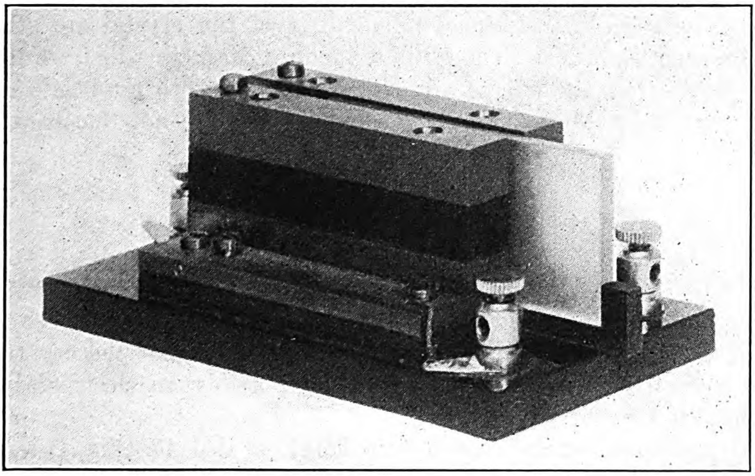

Fig. 3—Quartz Crystal in Flexural Vibration Mounting.
The crystal is at the left, projecting from the mounting to show the sled used to investigate the nature of translational movements.

distance traveled in either direction was found to be approximately the same. The oscillating crystal seems to have a tendency to tune the circuit to maximum power output by its motion as described above. The emergence of the crystal from the mounting diminishes the capacity of the system since a dielectric of air is then replacing the quartz dielectric in the crystal mounting. This means also that the circuit is operating at a different point on the characteristic curve of the crystal. A photograph of one of the crystals with flexural vibration mounting as used in these tests is shown in Fig. 3.

The same translational movements of the crystal are still observed when it is enclosed in a vacuum chamber which would indicate that the effects are probably not due to air blasts as described by Meissner.[8] If the movements are not due to air blasts the only probable explanation would be that the crystal creeps on its base when vibrating, or possibly an electrostatic effect in the dielectric.

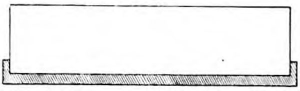

Fig. 4—Quartz Crystal Standing on Sled (cross-hatched) Which Prevented the Creeping of the Crystal When Oscillating.

To determine the cause a very light fiber sled *AB* (Fig. 4) was constructed to fit the crystal. The ends of the sled were made projecting as shown in the same figure so the crystal could not creep off. The crystal is also shown on the sled and in the mounting by Fig. 3. If the motion was due to a creeping effect of the vibrating crystal against its mounting the sled will now arrest the motion. If the motion was due to a dielectric phenomenon it should still take place with the sled, provided the weight added was negligible and the frictional effects were still small. Care was taken to reduce the friction between the sled and the base of the crystal mounting to a minimum by careful polishing and lubrication. No tendency of the crystal to move was noted when the sled was used although crystal oscillations were obtained just as before. Evidently, then, the motion is due to the vibration of the crystal against its support with a resulting creeping tendency.

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[8] A. Meissner, loc. cit.

PIEZO-ELECTRIC OSCILLATOR WITH TWO-ELECTRODE MOUNTING

Another new piezo-electric oscillator circuit is illustrated in Fig. 5. Here the crystal has a two-electrode mounting and is connected between the control grid and plate of the vacuum tube. This circuit is more stable than the crystal-oscillator circuits now commonly used, and at the lower radio frequencies gives a greater power output. As in the previous circuit a gridleak *R* is connected from the filament to the control grid, but for greater power output this is replaced by a choke coil and grid-bias battery. The plate circuit *LC* is tuned approximately to crystal frequency. The shield grid *S* as in the previous case is given a positive potential equal to about one-third the plate potential.

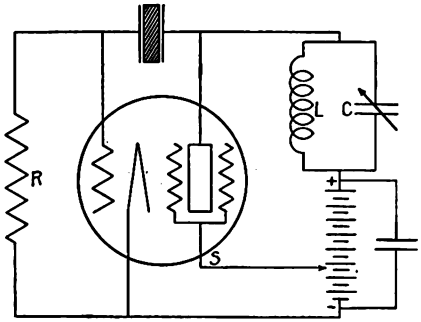

Fig. 5—Piezo-Oscillator with Screen-Grid Tube Similar in Operation to the Pierce Circuit.
The crystal is connected between the control grid and anode.

Also, as before, the shield-grid potential should be adjusted to the optimal value for maximum power output.

The four-electrode tube is here used as a shield-grid amplifier. The crystal does, however, annul the effects of shielding to a certain extent by introducing a capacity from the control grid to the plate of the vacuum tube.

This circuit (Fig. 5) is analogous to the piezo-electric oscillator using a three-electrode tube first described by Pierce.[9] This is the circuit in which the crystal is connected from grid to plate and is illustrated in Fig. 6. Oscillations are sustained in this circuit by energy feedback through the grid to filament inter-electrode capacity of the vacuum tube. It is well-known that this circuit is analogous to the Colpitts oscillator circuit with the crystal acting the part of an inductive reactance.

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[9] G. W. Pierce, *Proc.* Amer. Acad., 59, 81; 1923

From the practical point of view, the new circuit here described (Fig. 5) is more convenient than the circuit of Fig. 2, since the crystal mounting has but two electrodes like those now commonly used. The circuit possesses the disadvantage, however, that it is not as constant in frequency with variations in circuit constants as the circuit of Fig. 2.

The shielded-grid, four-electrode vacuum tube will not function as a piezo-electric oscillator with the crystal connected from control grid to filament. This is not possible, since in that type of circuit oscillations are sustained by energy feedback through the control grid to plate inter-electrode capacity and in the shield-grid tube this capacity is so small that the feedback is negligible.

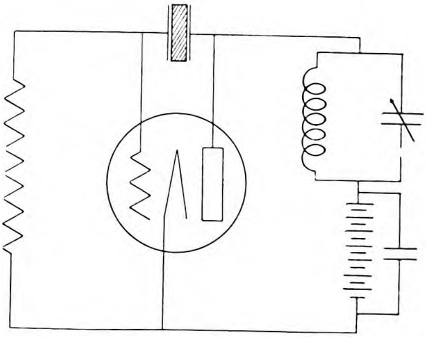

Fig. 6—Pierce Type of Piezo-Oscillator with Crystal Connected between Grid and Anode.

The circuit can be made to oscillate with the crystal connected from the control grid to filament if the shield grid and plate are connected together and used commonly as the anode of the system. There is no advantage in doing this, however, since we then have what is effectively a three-electrode tube oscillator as commonly employed.

HULL'S PIEZO-ELECTRIC OSCILLATOR CIRCUIT

A. W. Hull[10] has recently described still another piezo-electric oscillator using the four-electrode tube. Hull was the first to describe a piezo-electric oscillator circuit using the four-electrode tube. This circuit, which is illustrated in Fig. 7, uses the four-electrode valve as a space-charge-grid tube. The crystal Q is connected from the filament to the grid nearest the plate (now the control grid). The inner grid G (nearest the filament) now becomes the space-charge grid and is given a positive potential from the plate supply battery B. In Hull's patent the tuned circuit LC (Fig. 7) is replaced by an inductance L having a natural frequency at least twice that of the crystal used. This circuit is then coupled to a four-electrode tube (shield-grid) amplifier with a tuned output circuit. Hull claims extreme constancy of frequency for this circuit—"With a proper choice of circuit constants and the temperature of the piezo-electric element maintained constant, the generator remains constant to less than one part in a million." The power output of this type of circuit as shown in Fig. 7 does not compare with that obtained from the new circuits here described with the four-electrode tubes now available.

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[10] A. W. Hull, British Patent No. 266, 690; March, 1927.

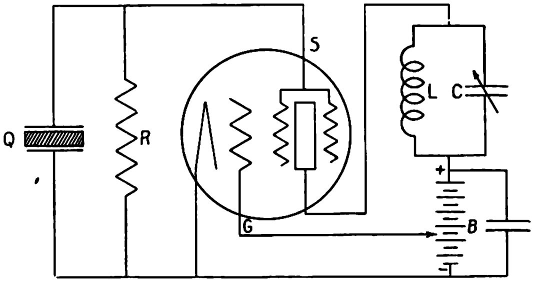

Fig. 7—Hull's Piezo-Oscillator Circuit Using the Four-Electrode Tube As A Space-Charge-Grid Device.

TWIN OSCILLATION FREQUENCIES

Experiments show that crystals will oscillate at two frequencies quite close together for each mode of vibration in the four-electrode vacuum-tube circuit of Fig. 2. This only happens, however, when the two pairs of crystal electrodes AC and BD are close together. Under certain conditions, then, oscillations are obtained at two different points on the same resonance curve of the crystal. When the two pairs of crystal electrodes are not close together only one oscillation frequency is found for each mode of crystal vibration as with the familiar three-electrode tube circuits.

Normally crystal oscillations are sustained in a circuit of this type by energy feedback through the vibrating crystal itself. This was found by experiment to be the lower of the two oscillation frequencies obtained, and the one which always remained when the pairs of crystal electrodes were moved far apart.

The higher of the two oscillation frequencies obtained was found by frequency measurements to be the same that is obtained from the crystal when it is connected simply between the control grid and the plate as in Fig. 3. In other words, this is the case where the crystal acts as an inductive reactance like the inductance in a Colpitts oscillator circuit. The measurements above referred to were simply frequency measurements of the two oscillation points of the circuit of Fig. 2 and the single oscillation point of the circuit of Fig. 3. For all of the three frequency measurements the same crystal was used on the same mode of vibration. The method of measurement was to obtain an audio-frequency beat note between the piezo-electric oscillator in question and another piezo-electric oscillator of nearly the same frequency. The beat note was amplified on a two-stage, audio-frequency amplifier and measured from the readings on a sonometer driven by a telephone receiver connected to the output of the amplifier.[11] For accurate determinations of frequency from the sonometer, reference is made to Allan's corrections to the well-known formula due to Brook Taylor.[12]

If the electrical connections to one pair of metallic electrodes such as *AC* (Fig. 2) are reversed, oscillations will still be maintained on the fundamental mode of vibration. Cady found with the three-electrode, vacuum-tube, piezo-electric oscillator of Fig. 1 that the connections to a pair of electrodes should be reversed in this way to obtain oscillations on the first harmonic of the transverse effect.[13] He did not, however, obtain oscillations on the fundamental with the reversed connection. The circuit now under consideration (Fig. 2) has with the reversed connection one oscillation frequency for the transverse fundamental and two oscillation frequencies close together for its first harmonic. The reversed connection then suppresses the lower oscillation frequency of the transverse fundamental. This would be expected since the proper phase relationship for sustaining oscillations by feedback through the crystal is not obtained with the reversed connection.

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[11] C. B. Jolliffe and G. Hazen, Bureau of Standards Scientific Papers, 21, 179; 1926.

[12] G. E. Allan, *Phil. Mag.*, 4, 1324; 1927.

[13] W. G. Cady, PROC. I. R. E., 10, 111; April, 1922.

Tests were then made to determine whether the lower of the two oscillation frequencies obtained is due to feedback through the crystal. A long crystal of 10.3×2.0×0.1 centimeters was secured which was fitted to a crystal mounting having two pairs of electrodes. The electrodes were adjustable so that the distance between the pairs of electrodes *AC* and *BD* (Fig. 2) could be varied so that they might be very close together or at the opposite ends of the crystal. When the pairs of crystal electrodes are separated by about one-half centimeter or more, oscillations are obtained only at the lower frequency corresponding to each mode of vibration. Also, when the electrodes are thus separated and the connections to one pair reversed, no oscillations are obtained at frequencies corresponding to the fundamental modes of the transverse and longitudinal effects. Oscillations are always obtained, however, with normal connections as in Fig. 2 at the lower oscillation frequency for the transverse and longitudinal fundamentals even though the pairs of electrodes are separated by 6 to 8 centimeters. This seems to indicate conclusively that the oscillations are sustained by feedback through the crystal.

FLEXURAL VIBRATIONS

Recently a method has been described for exciting very much lower frequencies than hitherto obtained from quartz plates with flexural vibrations in the length-breadth plane.[14] Considerable difficulty has been encountered, however, in making crystals oscillate at flexural vibration frequencies with a three-electrode tube circuit as commonly used. It was found necessary to use a pickup or sensitizing coil in series with the crystal in the grid circuit to obtain sufficient feedback to sustain oscillations. The only other alternative offered at that time was to connect the crystal to a three-stage, resistance-coupled amplifier using the method of feedback through the crystal. The energy feedback was then sufficient to sustain oscillations without the pickup coil.

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[14] J. R. Harrison, PROC. I. R. E., **15**, 1040; December, 1927. Author's note: More recently other investigators have described similar methods for obtaining still other flexural vibrations in various planes of the quartz plate. The investigations cover both the "Curie cut" and the "30-degree cut" quartz plates. A method of exciting torsional vibrations is also given. In all these cases, however, the crystal response is so feeble that it is only detected by the luminous glow surrounding the plates when they are vibrating in a partially evacuated chamber. For details see the following references:

E. Giebe and A. Scheibe, *Zeit. f. Instrumentkunde*, **47**, 269; 1927. *Zeit. f. Hochfrequenztechn.*, **30**, 32; 1927. *Zeit. f. Physik*, **46**, 607; 1928. *Elektr. Nachr.-Techn.*, **2**, 65; 1928.

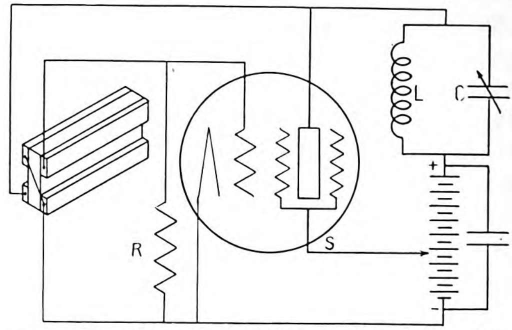

Fig. 8—Connections to the Flexural Vibration Mounting When Using the Screen-Grid Tube with Feedback through the Crystal.

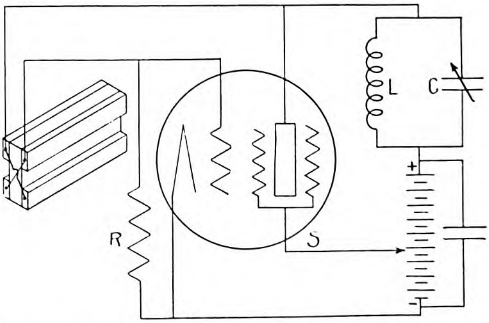

Fig. 9—Connections to Flexural Vibration Mounting When Using Screen-Grid Tube with Feedback through the Tube.

The use of a pickup coil in series with the crystal is a serious disadvantage since if the coupling is made too close the large induced voltages may puncture the crystal or set up parasitic oscillations independent of it. The inconvenience of the three-stage amplifier scheme is only too apparent and the power output is comparatively small.

The four-electrode tube, crystal-oscillator circuits are used to considerable advantage at low frequencies. With the four-electrode tubes now available, crystals oscillate at the flexural mode without a pickup coil. Either of the new circuits here described can be used. The four-electrode tube circuits can be used to equal advantage at the higher modes of flexural vibration.

The screen-grid tube oscillator circuit utilizing feedback through the crystal is illustrated in Fig. 8 as applied to a crystal for flexural vibrations. The circuit is exactly the same as that shown in Fig. 2 except for the crystal mountings and the connections thereto. Fig. 9 illustrates the application of the circuit of Fig. 5 to a crystal with a flexural vibration mounting. In this type of circuit, as was explained before, oscillations are sustained by feedback through the vacuum tube.

Finally, the author wishes to thank Professor W. G. Cady for his encouragement and many helpful suggestions during the progress of this work. Thanks are also due to Professor K. S. Van Dyke for suggestions.

Discussions

August Hund†: We have just listened to a most interesting contribution to this subject. No doubt with the advent of the two-grid tube everyone working with the piezo-crystal must have recognized in such devices a means for better stabilization than that found in the ordinary tube circuits. I was glad that Mr. Harrison brought out in his paper tonight a point which I distinctly missed in the preprint of his paper. The point to which I have reference has to do with his statement, "these circuits are more stable." In tonight's presentation the speaker has told us very clearly that by "stable" he means that the oscillations start more easily, or when stopped by some means come back more easily. He, therefore, makes a distinction between stability of this type and stability with respect to frequency. Usually we think of a stabilized crystal circuit as one the frequency of which is very constant. It may therefore be advisable in the final publication of the paper to bring this point out clearly in order to avoid any misunderstanding on the part of the reader.

† Radio Division, Bureau of Standards, Washington, D. C.

I do not agree that the circuit is stable with respect to frequency when it gives more power. I doubt that very much; I think it is the opposite. Any circuit that has very much back feed is apt to run away—to swing very much from side to side, but when its amplitude of vibration is very, very small, it is more apt to stay right on the correct frequency.

Another thing which has occurred to me which was not brought out in the paper is this: Mr. Harrison realized that when we use a shield-grid tube in the real sense of a shield-grid tube (with proper shielding voltage), we cannot connect the crystal between the filament and the grid and expect oscillations because the capacity—call it effective capacity—between the control grid and the plate may be only one per cent of the physical capacity which it would have without the plate shielded; but it will oscillate. All we have to do is to use the circuit as he had it. The paper described a circuit in which the crystal is connected between the grid and the plate, using the crystal capacity as a bridge to feed back. We can also connect the crystal between the control grid and the filament and use a very small condenser between the control grid and the plate. I think this is a better circuit than the one described in which the crystal is between the control grid and the plate, because we can use a very small air condenser of good construction and produce very stable oscillations. With this circuit you can produce all three oscillations very easily. In the Sound Laboratory of the Bureau I have used this circuit for some two years with much success.

J. R. Harrison: I had supposed that the context would make clear what I meant by stability—a circuit which starts oscillating very readily without retuning.

With regard to making a screen-grid tube oscillate with the crystal connected between control grid and anode, I would say that the interlectrode capacity of the tube which is always present in parallel with the crystal is now very much smaller than if a three-electrode tube had been used. The tube is not of course then used in its complete sense as a screen-grid tube, but we are still benefited but its screening action because of the very small capacity shunted across the crystal.

I have succeeded in making a four-electrode tube oscillate with a piezo-electric crystal connected between the control and filament. This cannot be done, however, with a four-electrode tube used as a screen-grid tube. If the screen-grid tube is used in this way it is necessary to introduce a capacity between control grid and anode as Dr. Hund has done, or to make the screen-grid potential very small. Both methods then producee the same result, i.e., a reduction of the screening action of the valve.

August Hund: Then you have no screen-grid tube any more.

J. R. Harrison: But it is doing just the same thing you did only in a slightly different way.

August Hund: I do not think it is the same because it is an easy matter to connect a small fixed condenser externally, whereas the interelectrode capacity (*effective* interelectrode capacity) does not seem very constant. Changes in the filament emission, potential variations on all electrodes, etc., no doubt make the capacity rather indefinite. I regard the external capacity method as being decidedly better.

J. R. Harrison: In the Cady type of piezo oscillator which uses mechanically tuned feedback through the crystal (Fig. 2) the pairs of crystal electrodes may be 8 or 10 centimeters apart and oscillations will still be maintained. The capacity introduced across the control grid and anode of the screen-grid tube is then very small. It would seem that this is the only method of applying the crystal to the screen-grid tube which does not appreciably reduce the screening action.

August Hund: There is another question that occurs to me. The paper brings out that one should adjust the screen-grid potential to an optimum—call it an energy optimum. Does that potential which you apply still keep the screen-grid tube screened, or have you changed it so much that you have a case of a feedback, and you get more energy on account of this?

J. R. Harrison: I feel quite sure it is still being used as a screen-grid tube when using the circuit of Fig. 2. The screen-grid potential is not very critical, but there is a definite optimum value for maximum power output. Under these conditions the screen-grid current is very small, usually less than one-tenth of the anode current, thereby indicating that the tube is virtually performing as a screen-grid tube.

W. G. Cady†: Regarding the two frequencies very close together which Mr. Harrison has described, at either one of which the circuit may be made to oscillate, a word may be said. You will recall that they occur only with a quartz plate having two pairs of electrodes, one pair being connected between filament and grid, and the other between filament and anode. The transition from one frequency to the other comes when the two pairs of electrodes are at a certain critical distance apart. What I wish to point out is that these two frequencies are not due to two different modes of vibration of the crystal, but that they simply represent two different operating points on the resonance curve relating amplitude of vibration to frequency. The point on the curve at which any crystal operates depends on circuit conditions. We have one set of circuit conditions when the electrodes are far apart, so that feedback is due solely to the mechanical vibrations of the quartz. On gradually moving the electrodes more closely together, the effective grid-anode capacity begins to be appreciable and increases until quite abruptly feedback begins to take place electrostatically instead of mechanically. This is the change in circuit conditions, I think, that accounts for the change in frequency.

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† Department of Physics, Wesleyan University, Middletown, Conn.

It is hardly necessary to add that this is quite a different phenomenon from the abrupt "jumps" in frequency that are sometimes so troublesome with crystal-controlled oscillators of the ordinary type. Such jumps are due to temperature changes and other causes, and represent, usually, a shift from one mode of vibration to another. They are only found, of course, when there happen to be two or more possible modes of vibration of nearly the same frequency.

Alfred N. Goldsmith†: It is indeed most fortunate that just when extremely accurate frequency control is becoming so necessary to radio, the instrumentalities for it should become available. It may be that this is putting the cart before the horse and explaining how remarkable it is that great rivers always flow past large cities. Perhaps the modern desire and need and application of constant frequency is the result of the crystal oscillator and similar high precision frequency control devices.

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† Chief Broadcast Engineer, Radio Corporation of America, New York City.

Volume 16, Number 11 *November, 1928*

NOTE ON THE DETERMINATION OF THE IONIZATION IN THE UPPER ATMOSPHERE*

By

J. C. SCHELLENG

(Bell Telephone Laboratories, New York City)

Summary—*The paper describes a method of estimating the distribution of ionization in the upper atmosphere. It is based upon measurements on several frequencies of the effective height as determined by interference or echo experiments. The latter two types of experiment are shown to give identical results.*

A NUMBER of ingenious radio experiments have been devised and carried out by which estimates of the distribution of ionization in the upper atmosphere have been obtained. Most of these lead to an effective height which is arrived at by assuming regular reflection. Plausible assumptions are sometimes invoked to permit the calculation of corrections which take account of the fact that the density of ionization varies continuously. These calculations usually involve the method of trial and error, and while some of them may possibly lead to fairly accurate results, a more direct method is desirable.

One of the purposes of the present note is to discuss a different method which may possess certain advantages. As a preliminary we will discuss certain relationships between some of the different methods which have been used.

In one class of experiments the time required for a pulse to travel by the overhead path to the receiver is measured. While in Breit's experiment this is done directly by time measurements we must also include in this class experiments such as those of Bown, Martin, and Potter; Appleton and Barnett; and Heising, in which by means of a slow shift of frequency, the number of "fringes," i.e., maxima of field intensities, is counted, either by observing a meter or by recording with an oscillograph. That this type of measurement gives the time required for a pulse to travel between the stations (group time) is not obvious. The mistake is sometimes made of assuming that the total number of wavelengths in the path can be obtained in this way. Thus Pedersen[1] states that this type of measurement gives values of

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* Original Manuscript Received by the Institute, August 27, 1928.

[1] "Propagation of Radio Waves," p. 229.

the height of the apex of the path which are too small, a statement which would be correct if the measurement gave the total number of wavelengths. As a matter of fact this type of measurement in general gives heights which are too great, as in the case of echo experiments.

This can be shown as follows. The experiment gives directly the quantity dN/df, where N is the total number of wavelengths (N_2) in the upper path minus the total number (N_1) in the direct path. In the case of reflection at a sharp surface of discontinuity we would have $N=D/\lambda=fD/c$ where D is the difference of the distances of D_1 and D_2 and c is the velocity *in vacuo*. In this case

$$\frac{dN}{df}=\frac{D}{c}\text{ and }D=c\frac{dN}{df}\cdot \tag{1}$$

The results of experiments have been interpreted in this way[2], the effective height, h, of the apex of the path being calculated by simple triangulation.

Now in general, regardless of the path followed,

$$N=N_2-N_1=f(T_{p2}-T_{p1}) \tag{2}$$

in which T_p refers to the *phase time*, or the time required for a crest of the wave to travel between stations along the path indicated by the subscript.

Therefore

$$D=c\frac{dN}{df}=c\left(\frac{d(fT_{p2})}{df}-\frac{d(fT_{p1})}{df}\right) \tag{3}$$

But $$\frac{d(fT_p)}{df}\text{ is the }\textit{group time.}$$

Hence the height is the same as would be measured at the time by an echo experiment which would measure this time lag directly. It should be noted that this conclusion does not involve any assumptions as to the mechanism of transmission. It is true in general regardless of the paths of the two waves.

One conclusion to be drawn is that we should be able to compare the results of these two methods.

Breit has shown that in the case of waves for which the effect of collisions and the effect of the earth's magnetic field may be neglected, this triangulation, in the case of pulse experiments,

[2] The authors referred to fully appreciated that the results give only apparent heights.

also gives the correct earth angle. At lower frequencies these assumptions are not strictly true. In the range of short waves they are correct to the extent that we can depend on the assumption that ionization is not a function of the horizontal coordinates.[3]

Hence the several methods for measuring the height by determining the pulse time, the earth angle, and shift of interference fringes, lead substantially to the same result for short waves. *None of them give the total number of waves in the path.*

The method to be described requires a means for obtaining the total number of waves in the path. This might be done experimentally in the following difficult manner. Starting at a very low value, the frequency would gradually be increased. At a receiver located one or two hundred kilometers away the fringes would be counted. That these fringes exist has been found by Hollingworth at 20 kc, by Bown, Martin, and Potter at 610 kc, by Appleton and Barnett at 750 kc, and by Heising at about 5000 kc. Each fringe would represent the gain of one wave in the overhead path as compared with the number, also increasing, in the direct path. This number, integrated from zero frequency to the frequency in which we are interested, would give the difference in wave numbers for the two paths. Practically, we would have to start within the radio range, so that an estimate of these quantities at the lower frequency would have to be made. This, in itself a small correction, could be calculated with sufficient accuracy to make the final error very small providing that the upper frequency limit is large compared to the lower. Thus if these limits were respectively 5000 and 20 kc the error would be a very small fraction of a per cent.

Having obtained the wave number we would use the construction suggested by Pedersen,[4] shown in Fig. 1. According to his calculations, the wave number in the path APQ is equal to AQ/λ_0 where λ_0 is the wavelength *in vacuo*. This is not strictly true, but the error is small providing that the initial earth angle is less than 60 deg. from the horizontal. This therefore furnishes a method of determining the height of the apex of the path.

However, this experimental procedure is prohibitive and unnecessary. As shown above,

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[3] It also assumes the validity of the method of rays. This method should be subjected to a more critical examination than any it has yet received.

[4] "Propagation of Radio Waves," p. 176.

$$T_{g2}-T_{g1}=\frac{d}{df}(N_2-N_1)$$

Therefore

$$N_2-N_1=\int_0^{f_0}(T_{g2}-T_{g1})df=\int_0^{f_0}\frac{d}{df}(N_2-N_1)df \qquad (4)$$

Hence

$$AQ-AF=\frac{\lambda_0(N_2-N_1)}{2}=\frac{c}{2f_0}\int_0^{f_0}(T_{g2}-T_{g1})df. \qquad (5)$$

As pointed out above, the integrand can be obtained from different types of experiments. Having determined $AQ-AF$ the solution for FQ is obvious.

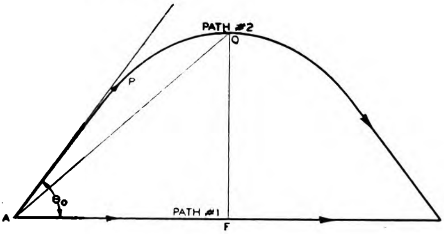

Fig. 1

We know that the group time at the lower frequencies is less than for the high. This method corrects the large apparent heights of the higher frequencies by means of the small apparent heights of the lower frequencies.

The next step is the calculation of the ionization on the basis of the earth angles calculated from the original data. Owing to complications due to the earth's magnetic field this can be done with confidence for the short waves only.

A further refinement in the estimate could be made by recalculating the number of waves in the path on the basis of the computed distribution of ionization. This would furnish a second approximation to the estimate for the height, the magnitude of the error being dependent on the earth angle and being small under the conditions already stated.

Data are not available for carrying out this calculation with confidence. It is of interest, however, to try the method with data which are based on actual experiment, but which are necessarily assumed to hold under somewhat different conditions from those which actually existed during the experiments.

In Fig. 2 the apparent night-time heights from the experiments mentioned are plotted on a kilocycle scale. The four points correspond to the results of Hollingworth, Bown, Martin and Potter; Appleton and Barnett; and Heising. It is rather interesting that these fall on a straight line, the equation of which is

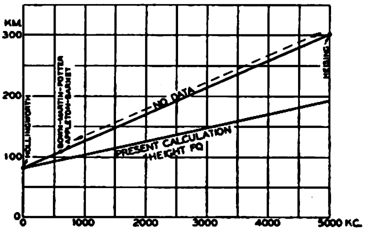

Fig. 2

$$H = 80 + 0.0440\,f \tag{6}$$

The exactness is accidental, however, since in the absence of knowledge regarding the time of year for some of these experiments we have averaged Hollingworth's summer and winter results. There is a very large part of the spectrum in which there are no data. The point for 5000 kc is the mean of widely different observed values.

Assuming for the sake of illustration that the base line ($2d$) had been 150 km in all these experiments, the difference in the group times along the two paths would have been

$$\frac{2}{c}(\sqrt{H^2+d^2}-d)$$

Hence by (5),

$$AQ - AF = \frac{1}{f_0}\int_0^{f_0}[\sqrt{H^2+d^2}-d]df \tag{7}$$

Substituting (6) in (7) and carrying out the indicated operations, the height FQ of the apex of the actual path can be calculated. This is shown in Fig. 2, which indicates that at the highest frequency shown the actual height for the conditions assumed would have been 192 km instead of 300. Owing to the relatively high angles implicitly assumed, the calculated figures are somewhat too low.

Fig. 3 gives the electrons per cubic centimeter calculated on the assumption that collisions and the earth's magnetic field do not greatly affect the results. These assumptions are satisfactory at frequencies higher than two or three megacycles but at lower frequencies the earth's magnetic field will produce errors.

We wish to emphasize that these calculations are given as an example of a method. While the estimate cannot have much weight on account of the meagerness of the data, the results nevertheless do look plausible, and are probably more accurate than the original data.

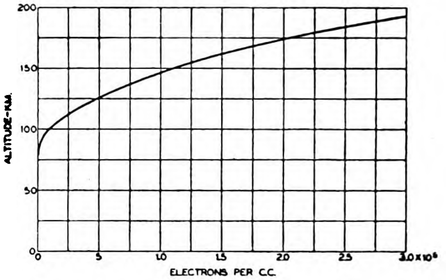

Fig. 3

CONCLUSION

The similarity between interference experiments and pulse experiments is pointed out. A method for obtaining the number of waves in the trajectory of the overhead wave is described. It is then possible to calculate the height of the apex of the path. From ray theory the number of electrons per cubic centimeter can then be calculated for certain conditions. By combining the results of different experimenters and making certain assumptions regarding the numerical quantities, the distribution is calculated, primarily as an example of the method. The results indicate an approximate increase in proportion to the second power of the height above 80 km, reaching a value of 3×10^5 at about 200 km. This is for night-time transmission.

The results indicate that a fruitful line of study will be to carry out echo or fringe experiments at several frequencies from 1000 to 10,000 kilocycles, the base line being made sufficiently large to avoid initial ray angles greater than 60 deg. from the horizontal. Such experiments should enable us to calculate the number of wavelengths in the path and hence the distribution of ionization. An experiment of this kind is now in preparation.

Volume 16, Number 11 *November, 1928*

ANALYSIS OF BROADCASTING STATION ALLOCATION*

By

J. H. Dellinger

(Chief Engineer, Federal Radio Commission, Washington, D. C.)

THE new allocation of broadcasting stations announced by the Federal Radio Commission on September 11, 1928 was prepared in accordance with the allocation plan set forth in the Commission's General Order No. 40, of September 7, 1928. Both the plan and the allocation itself were drawn in compliance with the requirements of the 1928 Amendment to the Radio Act as to equalization of broadcasting facilities between the zones and states. The allocation was, furthermore, made in compliance with the Commission's decision that no existing stations should be abolished at the time of its inception. It is believed to provide the greatest aggregate of radio service to the country possible under the two conditions just mentioned. Its principal features are: (a) it provides a definite, invariant basis of station assignments for each zone and locality, (b) it can be improved wherever interference is found to exist in actual operation, through the reduction of power or the elimination or particular stations, without disturbing the station allocation as a whole, (c) it eliminates heterodyne interference on 80 per cent of the listener's dial, (d) it recognizes the essentially different requirements of local, regional, and distant service.

Proper provision for the differing requirements of the listeners in large rural areas, cities, and intermediate areas made the preparation of this allocation a difficult task. It would have been very easy to allocate all existing stations, and many more, if only local service or the effects a few miles from the station had been considered. As soon as consideration was given to service more than a few miles from a station, serious difficulty arose, since heterodyne interference extends to many times the distance from a station to which actual program service extends. Operation of two or more stations on a channel (i.e., on one frequency or wavelength) results in an area of destructive interference very much greater than the area in which program service is provided unless the stations are of low power and widely spaced

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* **Original Manuscript Received by the Institute, September 15, 1928.**

geographically. It is only when a station has exclusive use of its channel that program service free from interference can be furnished at great distances. But since there are only 90 channels available for broadcasting in the United States, there could not possibly be more than 90 simultaneously operating stations giving service at great distances.

The only reasonable solution of this dilemma is that which the Commission has adopted, the setting aside of a certain number of channels (40) for distant or rural service, each with only one station assignment,[1] and the use of the remaining channels for service at more moderate distances with several station assignments on each channel, all with limited power and located systematically at proper distances apart to minimize interference.

The channels used for the latter type of station assignments are subdivided into "regional service" channels, which are kept substantially free from heterodyne interference by restricting power to 1000 watts and keeping the stations on a given channel in general 1000 miles or more apart, and several other types of channels on which heterodyne interference is permitted but which give satisfactory local service.

Besides the channels designated as "local service" there are two classes of "limited service" channels on which heterodyne interference is permitted. On five of these channels, 1000-watt stations are permitted, and on four of them 5-kilowatt stations. These will not give distant service and are in that sense "limited", but will give better local service than the stations on the "local service" channels because of their higher power. In some discussions the 1000-watt limited service channels are lumped with regional service channels, because there is not a very sharp difference between them; a heavily loaded regional service channel would be indistinguishable from a 1000-watt limited service channel.

There has been no specific designation of a name for the class of channels intended to give distant or rural service. They have been called variously "rural service," "distant service," "cleared," "high-power," "heterodyne-free," and "exclusive" channels. Stations on these channels may be authorized to use power up to 25 kilowatts, and, experimentally, up to 50 kilowatts.

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[1] The expression "station assignment," or "full-time assignment," indicates full-time operation 24 hours a day by a station, or a group of stations sharing time.

The allocation is in harmony with good engineering principles. In the separate provision for high-power exclusive channels and restricted-power local channels, and in the geographical spacings of stations on the same and adjacent frequencies, and in other vital respects, the allocation is in accord with "A Statement on Engineering Principles" presented to the Commission on March 30, 1927, by the Committee on Radio Broadcasting of the American Engineering Council. It is also in essential accord with the recommendations of the radio engineers in the April 6, 1928 conference, except that only 40 high-power exclusive channels are provided instead of 50.

SUMMARY OF ALLOCATION PLAN

The allocation plan is set forth in detail in General Order No. 40. Its principal features are indicated in the following table. The available numbers of station assignments have not in all cases been utilized in all the zones, in the allocation which the Commission has announced.

	High-Power, 5 kw and up	Regional, 500-1000 w	Limited Service		Local 10-100 w	Total
			5 kw	1000 w		
Number of Channels	40	35	4	5	6	90
Station Assignments per channel	1	[2] 2½	2½	5	25	—
Number Station Assignments in U. S.	40	90	10	25	150	315
Number Station Assignments in each zone	8	18	2	5	30	63

The allocation is based on night-time transmission conditions. Besides the classes of stations shown in the table, there are a number of supplementary stations added on some channels. These include a number of "daytime service" stations and "limited time" stations. The latter are allowed to operate during the day and also during certain time (after late evening in the East by western stations) temporarily not used by the stations entitled to the channel. The "day-time service" stations are allowed to operate only during non-interfering hours. They are required to shut down at sunset. This shall be taken to be sunset at the daytime service station unless it is the farthest east of the stations on the channel, in which case sunset at the next station west on the same channel. The time of sunset varies from about 4:30 in December to 7:30 in June, local sun time.

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[2] Approximate Average.

THE LISTENER'S DIAL

The choice of particular frequencies for the several classes of stations was influenced in considerable measure by the present frequencies of stations. Thus, one reason that the high-power channels are begun at 640 kilocycles rather than at 550 kilocycles is because the public is accustomed to hearing some of the regional service stations at this end of the spectrum. This principle has permitted reducing as much as possible the average shift of frequency which the stations must make.

The placing of several blocks of regional and local service channels in different parts of the dial has the advantage that it permits the licensing of more stations in certain places (e.g., Boston and Los Angeles) than would be possible (because of inter-channel interference) if the channels of each class of station were all bunched in a single group.

The high-power channels, however, are consolidated into a single block in the spectrum (except for Canadian exclusive and Canadian shared channels and the group of regional channels 880 to 950 kcs), so that the listeners on these heterodyne-free channels will be as free as possible from inter-channel interference from nearby stations of other classes.

The choice of channel locations is expected to have the effect of making programs as available at the high-frequency end of the listener's dial as at the low-frequency end. Thus the entire dial becomes useful, for listeners everywhere in the United States.

In the following list, the numbers in parentheses after certain frequencies indicate the zone to which that frequency is assigned.

550	Limited Service 1000
560	
570	
580	Regional Service
590	
600	
610	
620	
630	
640 (5)	Rural Service (i.e., high power)
650 (3)	
660 (1)	
670 (4)	
680 (5)	
690	Canada
700 (2)	Rural Service (i.e., high power)
710 (1)	
720 (4)	
730	Canada

740 (3)	Rural Service (i.e., high power)
750 (2)	
760 (1)	
770 (4)	
780	Regional Service (shared with Canada)
790 (5)	Rural Service (i.e., high power)
800 (3)	
810 (4)	
820 (2)	
830 (5)	
840	Canada
850 (3)	Rural Service (i.e., high power)
860 (1)	
870 (4)	
880	Regional Service
890	
900	
910	Canada
920	Regional Service
930	
940	
950	
960	Canada
970 (5)	Rural Service (i.e., high power)
980 (2)	
990 (1)	
1000 (4)	
1010	Regional Service (shared with Canada)
1020 (2)	Rural Service (i.e., high power)
1030	Canada
1040 (3)	Rural Service (i.e., high power)
1050 (5)	
1060 (1)	
1070 (2)	
1080 (3)	
1090 (4)	
1100 (1)	
1110 (2)	
1120	Regional Service (shared with Canada)
1130 (5)	Rural Service (i.e., high power)
1140 (3)	
1150 (1)	
1160 (4)	
1170 (2)	
1180 (4)	
1190 (3)	
1200	Local Service
1210	
1220	Regional Service
1230	
1240	
1250	
1260	
1270	
1280	
1290	
1300	
1310	Local Service
1320	Regional Service
1330	
1340	
1350	
1360	
1370	Local Service

1380	Regional Service
1390	
1400	
1410	
1420	Local Service
1430	Regional Service
1440	Limited Service 1000 w
1450	
1460	Limited Service 5 kw
1470	
1480	
1490	
1500	Local Service

EQUALIZATION

The table given above under "Summary of Allocation Plan" shows how the frequencies are equalized between the zones. Each zone receives exactly one-fifth of the station assignments. In some zones there are a few vacancies in the station assignments which will be available until future stations are constructed in the localities where those station assignments can be used. The allocation of frequencies and of station assignments to the individual states is closely proportional to population, as the law requires; this correspondence, of course, cannot be exact because the inequalities of state populations lead to many fractional quotas.

The aggregate power assigned to the stations is nearly equal for the five zones and is closely proportional to the populations of the states within each zone. For the future, moreover, the potential power of stations is exactly equalized between the zones, since by General Orders 40 and 42 the same upper limit of power is prescribed for all stations of each class.

The number of licenses is equalized only approximately, as follows: Zone No. 1, 108; Zone No. 2, 106; Zone No. 3, 115; Zone No. 4, 155; Zone No. 5, 132. The total number of licenses, or stations, is 616, an average per zone of 123. The principal disparity is an excess of 32 over the average, in the Fourth Zone (the Middle West). These departures from equality are inherent in the Commission's fundamental decision that no existing stations should be abolished at the time of the inception of the new allocation.

The equalization of time "on the air" is indicated essentially by the distribution of "station assignments," which is equal as between the zones, and reasonably proportional to population as between the states. The equalization of time is somewhat altered, however, by the addition of "daytime service" stations on some of the channels.

CONCLUSION

The channels are carefully cleared of inter-channel interference in every part of the dial. This clearing is particularly well effected in Zones 3, 4, and 5. Zones 1 and 2 being smaller, the geographical spacings are somewhat less than in the other zones, and interference may in a few cases be perceptible on winter nights.

It is believed that heterodyne interference is substantially eliminated except on the 9 limited service channels and the 6 local service channels. If such interference should develop on any of the 75 heterodyne-free channels, the Commission may reduce it by decreasing a station's power or eliminating one or more stations.

The principal features of the allocation, such as the assignment of amounts of power and of particular frequencies to particular localities, can not in general be altered, because of the interdependence of the frequency and distance separations throughout the entire set-up. However, the selection of stations in a given locality to be put in a particular power class, the selection of stations in a locality to be assigned to the specific frequencies allotted to the locality, and the relative amounts of time divisions by groups or stations, are all features which can be changed at any time as the Commission sees fit without affecting the soundness of the set-up in any way. Thus the Commission will have a quick and definite way of determining what its action should be on all broadcast license applications.

STATE QUOTAS OF BROADCASTING ASSIGNMENTS

General Orders Nos. 40, 41, and 42, of the Federal Radio Commission, published in the Radio Service Bulletin for August 31, 1928, outline the basis for a general reallocation of the broadcasting stations of the United States.

In effect, the Commission's orders recognize three principal classes of stations and specify the broadcasting channels which shall be used by each class. The existing stations are then assigned to channels in accordance with this plan, time divisions being required where necessary, in order to minimize interference and to make the apportionment of full-time assignments as required by the law.

		State Quotas of Full-time "Assignments"		
	Percentage of Total National Facilities Due State	1 "Rural" Service 5-kw and above	2 "Regional" Service Chiefly 500–1000-w	3 "Local" Chiefly 50-w and 100-w
SECOND ZONE				
Pennsylvania	7.0	2.80	8.8	10.5
Virginia	1.8	0.72	2.3	2.7
West Virginia	1.2	0.48	1.5	1.8
Ohio	4.9	1.96	6.1	7.3
Michigan	3.3	1.32	4.1	4.9
Kentucky	1.8	0.72	2.3	2.7
	20. per cent	8.	25.	30.
THIRD ZONE				
North Carolina	2.1	0.84	2.6	3.1
South Carolina	1.3	0.52	1.7	2.0
Georgia	2.3	0.92	2.9	3.4
Florida	1.0	0.40	1.3	1.5
Alabama	1.8	0.72	2.3	2.7
Tennessee	1.8	0.72	2.2	2.7
Mississippi	1.3	0.52	1.6	1.9
Arkansas	1.4	0.56	1.7	2.1
Louisiana	1.4	0.56	1.8	2.1
Texas	3.9	1.56	4.9	5.9
Oklahoma	1.7	0.68	2.2	2.6
	20. per cent	8.	25.	30.
FOURTH ZONE				
Indiana	2.4	0.96	3.0	3.6
Illinois	5.5	2.20	7.0	8.3
Wisconsin	2.2	0.88	2.8	3.3
North Dakota	0.5	0.20	0.6	0.7
Minnesota	2.0	0.80	2.5	3.0
South Dakota	0.5	0.20	0.7	0.8
Iowa	1.8	0.72	2.3	2.7
Nebraska	1.1	0.44	1.3	1.6
Kansas	1.4	0.56	1.7	2.0
Missouri	2.6	1.08	3.3	4.0
	20. per cent	8.	25.	30.
FIFTH ZONE				
Montana	1.0	0.40	1.2	1.5
Idaho	1.0	0.40	1.2	1.4
Wyoming	0.4	0.16	0.5	0.7
Colorado	2.0	0.80	2.4	2.9
New Mexico	0.7	0.28	0.9	1.0
Arizona	0.8	0.32	1.0	1.2
Utah	0.9	0.36	1.2	1.4
Nevada	0.1	0.04	0.2	0.2
Washington	2.8	1.14	3.5	4.2
Oregon	1.6	0.64	2.0	2.4
California	8.2	3.28	10.2	12.1
Ter. of Hawaii	0.5	—	—	—
Alaska	0.1	—	—	—
	20. per cent	8.	25.	30.

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THE DEPENDENCE OF THE FREQUENCY OF QUARTZ PIEZO-ELECTRIC OSCILLATORS UPON CIRCUIT CONSTANTS*

By

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Summary—*The mathematical theory for the quartz piezo-electric stabilized, vacuum-tube-driven oscillator is given for the following cases: tuned plate circuit, inductance-loaded and resistance-loaded triode with the crystal between grid and plate, and also between grid and filament for each case. The condition for oscillations and the exact expression for the frequencies, damping factors, coupling coefficient, tube constants, etc., is given. In the analysis of the oscillator the equivalent network for the crystal given by Van Dyke has been used. The theory has been checked by measuring the variation in frequency of a quartz stabilized oscillator for variations in impedance of the plate circuit for the tuned circuit and resistance-loaded tube respectively. To satisfy the condition for oscillation it is necessary to use values for the equivalent resistance of the crystal somewhat smaller than those given by Van Dyke's formula. A discussion of the general method by which conditions for oscillation and expressions for the driven frequency of an oscillator may be obtained from the coefficients of differential equations up to the fourth order is included.*

THE use of quartz piezo-electric oscillators, as first described by Cady,[1] for stabilizing the frequency of triode-driven circuits, has been of inestimable value in the maintenance of frequency standards and in holding radio transmitters on their assigned frequencies. Investigation has shown that the frequencies of such oscillators are, however, subject to certain variations, the chief causes of which are changes in temperature of the quartz plate, methods of mounting it, and the reactions of the electric circuit upon the mechanical properties of the quartz itself. By use of suitable thermostatic devices and by standardization of mountings, the first two sources of error have been brought under very good control, but the extent to which the frequency is influenced by the elements of the system of which the crystal forms a part has not as yet been fully investigated, and it is to supply this need in certain of the more

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* Original Manuscript Received by the Institute, August 6, 1928. Presented before the International Union of Scientific Radiotelegraphy, Washington, D.C., April 19, 1928.

[1] Cady, W. G., "Piezo-electric Resonator," PROC. I.R.E., **10**, **83**; April, 1922.

commonly used circuits that the following work has been carried out.

Van Dyke[2] has shown that a piezo-electric quartz crystal is equivalent electrically to an inductance, a capacitance, and a resistance joined in series, as shown in Fig. 1, with a second capacitance shunted across them, and has given formulas by which these equivalent electrical quantities may be computed for any of its normal modes of oscillation from the dimensions of the crystal. Using this method of representation, Dye[3] has studied the effect of the constants of the circuit to which the crystal is connected when used as a resonator, but as far as the author is aware, no study has yet been made of the corresponding problem—namely, the effect of the constants of the circuit when the crystal is used as a stabilizer for a vacuum-tube-driven circuit.

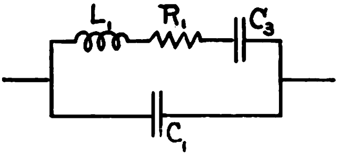

Fig. 1—Equivalent Network.

In carrying out this work, the crystal has been replaced in the circuits to be studied by its electrical equivalent as given by Van Dyke. It thus forms one of the elements of a coupled system, and the resultant frequency of the driven coupled system has been worked out in terms of that of the equivalent crystal element when uncoupled and oscillating according to one of its normal modes. In making the analysis of the various circuits considered, the method of differential equations rather than complex algebra has been used, since it permits the determination of the condition for oscillation and the calculation of frequency to be made directly from the coefficients of the differential equation. The advantages of this method, which, of course, are well-known, seem to be insufficiently appreciated, and a brief discussion of the theory may not be out of place.

Suppose we have any system, mechancial or electrical, such that its instantaneous state of motion may be described by a differential equation of the second order, e.g.,

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[2] Van Dyke, K. S., "The Piezo-electric Resonator and its Equivalent Network," Proc. I.R.E. 16, 742; June, 1928.

[3] Dye, D. W., "Piezo-electric Quartz Resonator and Equivalent Electrical Circuit," *Proc.* Phys. Soc. London, 38, 399; 1926.

$$\frac{d^2x}{dt^2}+P_1\frac{dx}{dt}+P_2x=0. \tag{1}$$

In order that $x=e^{mt}$ may be a solution of (1), it is necessary that the auxiliary equation

$$m^2+P_1m+P_2=0 \tag{2}$$

be satisfied—that is

$$m=\frac{-P_1\pm\sqrt{P_1^2-4P_2}}{2} \tag{3}$$

The condition that the system may execute oscillations is that $P_1^2<4P_2$. If this condition is satisfied, then the roots of (2) may be written

$$\left.\begin{array}{lll} m_1=\alpha+j\beta & \text{where} \quad \alpha=-\dfrac{P_1}{2} \text{ and } \beta=\dfrac{\sqrt{4P_2-P_1^2}}{2}. \\ m_2=\alpha-j\beta & \qquad\qquad j=\sqrt{-1} \end{array}\right\} \tag{4}$$

The solution may then be written

$$x=Ae^{\alpha t}\sin\beta t. \tag{5}$$

The damping factor α is seen to be one-half the coefficient of the first order derivative in (1). In general, P_1 is positive and oscillations die out logarithmically. If, however, the system contains some source of energy or a regenerative device, P_1 may be zero, or even negative. In the former case, oscillations, once started, persist with constant amplitude, and in the latter they are built up. The condition, then, that oscillations may persist in any vacuum-tube-driven circuit whose instantaneous state may be described by a second order differential equation is that the coefficient of the first order derivative be zero. It is also to be noted by (4) that when this condition has been satisfied, the square of the radian frequency of the system is given by P_2, the coefficient of the absolute term.

Unfortunately, not many triode circuits can be described by a second order differential equation. The method may, however, be extended to cases requiring third and fourth order differential equations for their description. For example, suppose a third order equation is required, e.g.,

$$\frac{d^3x}{dt^3}+P_1\frac{d^2x}{dt^2}+P_2\frac{dx}{dt}+P_3x=0. \tag{6}$$

The auxiliary equation is then

$$m^3+P_1m^2+P_2m+P_3=0 \tag{7}$$

If the system, thus described, is to be capable of oscillations, two of the roots of (7) must be a pair of conjugate complexes, and the third, real. Let them be

$$\begin{aligned} m_1&=\kappa \\ m_2&=\alpha+j\beta \\ m_3&=\alpha-j\beta \end{aligned} \tag{8}$$

where α and β are the damping factor and radian frequency, respectively, and κ the reciprocal of the time constant for the non-oscillatory transient.

The condition that oscillations once started may persist, that is, $\alpha=0$, may be obtained from the theorem giving the relation between roots and coefficients in an algebraic equation. Thus

$$\begin{aligned} m_1+m_2+m_3&=-P_1 \\ m_1m_2+m_2m_3+m_3m_1&=+P_2 \\ m_1m_2m_3&=-P_3 \end{aligned} \tag{9}$$

Substituting (8) in (9) we have

$$\begin{aligned} \kappa+2\alpha&=-P_1 \\ 2\kappa\alpha+\alpha^2+\beta^2&=P_2 \\ \kappa(\alpha^2+\beta^2)&=-P_3 \end{aligned} \tag{10}$$

If $\alpha=0$, these become

$$\begin{aligned} \kappa&=-P_1 \\ \beta^2&=P_2 \\ \kappa\beta^2&=-P_3 \end{aligned} \tag{11}$$

Whence $\beta^2=P_2$ and $\beta^2=\dfrac{P_3}{P_1}$ (12)

The condition, then, that $\alpha=0$ is that these two values of β must be the same; that is

$$P_2 = \frac{P_3}{P_1} \tag{13}$$

Thus the condition for undamped oscillations and the expression for the frequency may be obtained from the differential equation without solving it.

For the case of the fourth order differential equation the procedure is similar, and, although the conditions are somewhat more complicated, the method is readily workable. Let the differential equation be

$$\frac{d^4x}{dt^4} + P_1\frac{d^3x}{dt^3} + P_2\frac{d^2x}{dt^2} + P_3\frac{dx}{dt} + P_4x = 0 \tag{14}$$

with its auxiliary equation

$$m^4 + P_1m^3 + P_2m^2 + P_3m + P_4 = 0. \tag{15}$$

The system which this equation describes may be either simply or doubly periodic. In the former case, (15) has two real and one pair of conjugate complex roots, and in the latter, two pairs of conjugate complex roots. For the doubly periodic case the roots may be written

$$\begin{aligned} m_1 &= \alpha_1 + j\beta_1 \qquad & m_3 &= \alpha_2 + j\beta_2 \\ m_2 &= \alpha_1 - j\beta_1 \qquad & m_4 &= \alpha_2 - j\beta_2 \end{aligned} \tag{16}$$

where α_1, α_2 and β_1, β_2 are the damping factors and radian frequencies, respectively, for the resultant oscillations. Again making use of the relation between roots and coefficients we have

$$\begin{aligned} 2(\alpha_1 + \alpha_2) &= -P_1 \\ \alpha_1{}^2 + \beta_1{}^2 + 4\alpha_1\alpha_2 + \alpha_2{}^2 + \beta_2{}^2 &= P_2 \\ 2\alpha_1(\alpha_2{}^2 + \beta_2{}^2) + 2\alpha_2(\alpha_1{}^2 + \beta_1{}^2) &= -P_3 \\ (\alpha_1{}^2 + \beta_1{}^2)(\alpha_2{}^2 + \beta_2{}^2) &= P_4 \end{aligned} \tag{17}$$

The imaginary terms drop out, since the coefficients of (15) are all real. We seek now the condition that one of the damping factors, e.g., α_1, may be zero. Putting $\alpha_1 = 0$ in (17) α_2 and β_2 may be eliminated, and there results

$$\beta_1{}^2 = \frac{P_3}{P_1} \text{ and } \beta_1{}^2 = \frac{P_2 \pm \sqrt{P_2{}^2 - 4P_4}}{2}. \tag{18}$$

The condition for undamped oscillations is then either

$$\begin{cases} \dfrac{P_3}{P_1}=\dfrac{P_2+\sqrt{P_2^2-4P_4}}{2} \\ \text{or } \dfrac{P_3}{P_1}=\dfrac{P_2-\sqrt{P_2^2-4P_4}}{2} \end{cases} \tag{19}$$

Since equations (17) are symmetrical with respect to the α's and β's, the same result would have been obtained had we sought the condition that $\alpha_2=0$. It may be shown, however, that in (19) the plus sign gives the condition for one damping factor zero, and the minus sign that for the other zero.

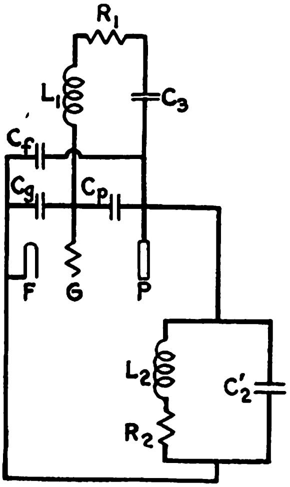

Fig. 2—Network for Crystal between Grid and Plate.

In stabilizing a vacuum-tube-driven circuit, the crystal may be connected either across the grid and filament of the tube or across the grid and plate. In the former case, the internal capacity between the plate and grid furnishes the necessary feed-back coupling, and in the latter, the grid-filament capacity serves this purpose. In this report, the solution will be given for these two cases using in each a tuned resonance circuit in series with the plate. The simplified ciruits of Pierce[4] may be discussed as special cases.

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[4] Pierce, G. W., "Piezo-electric Crystal Resonators and Crystal Oscillators Applied to Precision Calibration of Wave Meters." *Proc.* Am. Acad., 59, 81; 1923.

CRYSTAL BETWEEN GRID AND PLATE

I. Tuned Plate Circuit. The circuit diagram for this case is shown in Fig. 2, in which L_1, R_1, C_3 represent the equivalent series elements of the crystals as shown in Fig. 1, while C_p is the sum of the interelectrode capacitance between plate and grid of the tube and C_1 of the crystal. The simplified diagram is shown in Fig. 3, where $C_2 = C'_2 + C_f$. In setting up the Kirchhoff equations, the d.c. grid current has been neglected and a linear static tube characteristic has been assumed. With these simplifications, we have

$$L_1 \frac{di_1}{dt} + R_1 i_1 + \frac{1}{C_3} \int i_1 dt + \frac{1}{C_p} \int (i_1 - i) dt = 0 \tag{20}$$

$$L_2 \frac{di_2}{dt} + R_2 i_2 + \frac{1}{C_2} \int (i_2 - i - i_p) dt = 0 \tag{21}$$

$$\frac{1}{C_g} \int i dt - \frac{1}{C_2} \int (i_2 - i - i_p) dt - \frac{1}{C_p} \int (i_1 - i) dt = 0 \tag{22}$$

$$i_p = \frac{1}{R_p} (e_p + \mu e_g) \tag{23}$$

where $$e_p = \frac{1}{C_2} \int (i_2 - i - i_p) dt; \quad e_g = \frac{1}{C_g} \int i dt$$

Differentiating and combining, these become

$$L_1 \frac{d^2 i_1}{dt^2} + R_1 \frac{di_1}{dt} + \left(\frac{1}{C_3} + \frac{1}{C_p} \right) i_1 - \frac{1}{C_p} i = 0 \tag{24}$$

$$L_2 \frac{d^2 i_2}{dt^2} + R_2 \frac{di_2}{dt} + \frac{1}{C_2} (i_2 - i - i_p) = 0 \tag{25}$$

$$i = \frac{C_0}{C_p} i_1 + \frac{C_0}{C_2} (i_2 - i_p) \tag{26}$$

$$\frac{di_p}{dt} = \frac{1}{R_p} \left[\frac{1}{C_2} (i_2 - i - i_p) + \frac{\mu}{C_g} i \right] \tag{27}$$

where $$\frac{1}{C_0} = \frac{1}{C_p} + \frac{1}{C_g} + \frac{1}{C_2}$$

Substituting (26) in (24), (25), and (27), respectively, we have

$$i_2-i_p=L_1C_m\frac{d^2i_1}{dt^2}+R_1C_m\frac{di_1}{dt}+\frac{C_m}{C_a}i_1 \tag{28}$$

$$L_2\frac{d^2i_2}{dt^2}+R_2\frac{di_2}{dt}+\frac{1}{C_b}(i_2-i_p)-\frac{1}{C_m}i_1=0 \tag{29}$$

$$\frac{di_p}{dt}=\frac{1}{R_p}\left[\frac{1}{C_b'}(i_2-i_p)-\frac{1}{C_m'}i_1\right] \tag{30}$$

where

$$\frac{1}{C_a}=\frac{1}{C_3}+\frac{1}{C_d}; \qquad \frac{1}{C_d}=\frac{1}{C_p}-\frac{C_0}{C_p^{\,2}}; \qquad \frac{1}{C_b}=\frac{1}{C_2}-\frac{C_0}{C_2^{\,2}}$$

$$\frac{1}{C_b'}=\frac{1}{C_b}+\mu\frac{C_0}{C_2C_g}; \quad \frac{1}{C_m'}=\frac{1}{C_m}-\mu\frac{C_0}{C_pC_g}; \quad \frac{1}{C_m}=\frac{C_0}{C_2C_p}$$

C_a is the total capacitance of the primary circuit between the points *da*; C_d that between the points *dd*; and C_b that between the points *ee*.

$$C_d=C_p+\frac{C_2C_g}{C_2+C_g}; \quad C_b=C_2+\frac{C_pC_g}{C_p+C_g}$$

Substituting (28) in (29) and (30), respectively, there results

$$L_2\frac{d^2i_2}{dt^2}+R_2\frac{di_2}{dt}+\frac{L_1C_m}{C_b}\,\frac{d^2i_1}{dt^2}\,\frac{R_1C_m}{C_b}\,\frac{di_1}{dt}+\left(\frac{C_m}{C_aC_b}-\frac{1}{C_m}\right)i_1=0 \tag{31}$$

$$\frac{di_p}{dt}=\frac{L_1C_m}{R_pC_b'}\,\frac{d^2i_1}{dt^2}+\frac{R_1C_m}{R_pC_b'}\,\frac{di_1}{dt}+\frac{1}{R_p}\left(\frac{C_m}{C_aC_b'}-\frac{1}{C_m'}\right)i_1 \tag{32}$$

Differentiating (28) and substituting (32), we have

$$\begin{aligned}\frac{di^2}{dt}=L_1C_m\frac{d^3i_1}{dt^3}+\left(R_1C_m+\frac{L_1C_m}{R_pC_b'}\right)\frac{d^2i_1}{dt^2}\\ +\left(\frac{C_m}{C_a}+\frac{R_1C_m}{R_pC_b'}\right)\frac{di_1}{dt}+\frac{1}{R_p}\left(\frac{C_m}{C_aC_b'}-\frac{1}{C_m'}\right)i_1\end{aligned} \tag{33}$$

Substituting (33) and its derivatives in (31) the following fourth order differential equation results:

$$\frac{d^4i_1}{dt^4}+P_1\frac{d^3i_1}{dt^3}+P_2\frac{d^2i_1}{dt^2}+P_3\frac{di_1}{dt}+P_4i_1=0 \tag{34}$$

in which

$$\left.\begin{aligned}
P_1&=\frac{R_1}{L_1}+\frac{R_2}{L_2}+\frac{1}{R_pC_b'}\\
P_2&=\frac{1}{L_1C_a}+\frac{R_1R_2}{L_1L_2}+\frac{1}{L_2C_b}+\frac{1}{R_p}\left(\frac{R_1}{L_1}+\frac{R_2}{L_2}\right)\frac{1}{C_b'}\\
P_3&=\frac{R_2}{L_1L_2C_a}+\frac{R_1}{L_1L_2C_b}+\frac{1}{R_p}\left(\frac{1}{L_1C_aC_b'}+\frac{R_1R_2}{L_1L_2C_b'}-\frac{1}{L_1C_mC_m'}\right)\\
P_4&=\frac{1}{L_1L_2C_aC_b}-\frac{1}{L_1L_2C_m^2}+\frac{R_2}{R_p}\left(\frac{1}{L_1L_2C_aC_b'}-\frac{1}{L_1L_2C_mC_m'}\right)
\end{aligned}\right\} \quad (35)$$

Introducing now the uncoupled damping factors α_a and α_b, the uncoupled, undamped frequencies β_a and β_b, and the coupling coefficient τ, defined as follows:

$$\alpha_a=\frac{R_1}{2L_1}\qquad \beta_a^2=\frac{1}{L_1C_a}$$

$$\alpha_b=\frac{R_2}{2L_2}\qquad \beta_b^2=\frac{1}{L_2C_b}\qquad \tau^2=\frac{\dfrac{1}{C_m^2}}{\dfrac{1}{C_a}\cdot\dfrac{1}{C_b}}=\frac{C_aC_b}{C_m^2}$$

the coefficients of (35) become

$$\left.\begin{aligned}
P_1&=2(\alpha_a+\alpha_b)+\frac{1}{R_pC_b'}\\
P_2&=\beta_a^2+4\alpha_a\alpha_b+\beta_b^2+\frac{1}{R_p}(\alpha_a+\alpha_b)\frac{2}{C_b'}\\
P_3&=2(\alpha_b\beta_a^2+\alpha_a\beta_b^2)+\frac{1}{R_p}\left[(\beta_a^2+4\alpha_a\alpha_b)\frac{1}{C_b'}-\frac{1}{L_1C_mC_m'}\right]\\
P_4&=\beta_a^2\beta_b^2\left[1-\tau^2+\frac{R_2}{R_p}\left(\frac{C_b}{C_b'}-\frac{C_m}{C_m'}\tau^2\right)\right]
\end{aligned}\right\} \quad (36)$$

Depending upon the coefficients $P_1\cdots P_4$, the solution of (34) may represent four exponentially damped transients, two such transients and a periodic function, or a doubly periodic function. In the circuits with which we are generally concerned in radio work, the last of these possibilities usually obtains. Neglecting for the moment the effect of the tube, the normal modes of oscillation of the coupled circuits (a) and (b), Fig. 3, consist of two currents in each circuit with frequencies and damping factors β_1 and α_1, β_2 and α_2, respectively. At one of these two frequencies the current in one circuit is nearly in phase with that in the other, while at the other frequency, the two currents are nearly opposite in phase. By suitably connecting the tube to the circuit, one of the normal modes of oscillation, β_1, for example, may supply voltages to grid and plate, respectively, nearly opposite in phase, and regeneration will occur for this frequency, while for the other frequency β_2, the grid and plate voltages are nearly in phase and absorption of energy by the tube rapidly damps this oscillation out. However, by an appropriate change in the connections, the situation may be reversed, and regeneration occur for the frequency β_2 and absorption for β_1. Because of these phase relations between the primary and secondary currents for the two normal modes of oscillation, it is not possible to have regeneration at both frequencies at the same time with a single tube. In the case of crystal-stabilized oscillators with tuned plate circuit, the grid- and plate-voltage relations are such that the system oscillates according to one only of its normal modes when the crystal is connected between grid and plate, and to the other only, when connected between grid and filament. The latter case is discussed in Section IV.

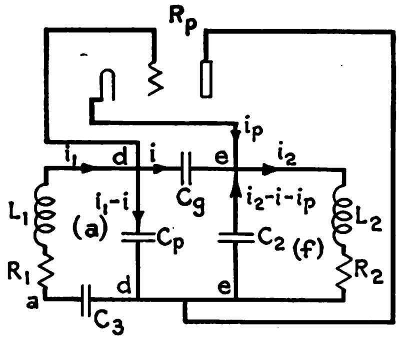

Fig. 3—Simplified Diagram for Fig. 2.

The action of the tube in driving such a circuit may be roughly regarded as neutralizing the resistance of the circuit for one of the frequencies, that is, by making one of the damping factors effectively zero, and at the same time increasing the other damping factor. In general, resistance neutralization is incomplete in that the frequency of the resulting oscillations is determined not by the inductances, capacitances, and degree of coupling of the circuits alone, but by the resistances of the two circuits and the electrical constants of the tube as well. The expression for the resultant frequency is given by (18), where where the P's of (36) are given such values that the condition for oscillation, (19), is satisfied.

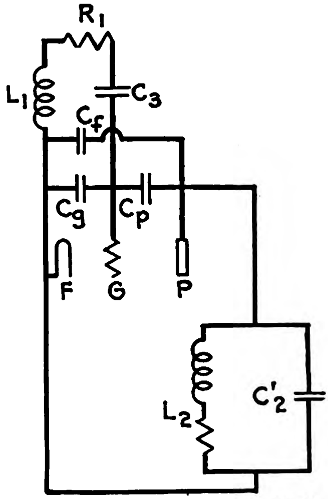

Fig. 4—Network for Crystal between Grid and Filament.

II. Inductance-Loaded Circuit. This may be regarded as a special case of the circuit just discussed in which C_2 is reduced to C_f, the capacitance between plate and filament of the tube. The circuit is still doubly periodic.

III. Resistance-Loaded Circuit. This is also a special case of the above in which, in addition to putting C_2 equal to C_f, L_2 is made zero. Thus, multiplying (34) by L_2 and then placing $L_2=0$ there results

$$\frac{d^3i_1}{dt^3}+P_1\frac{d^2i_1}{dt^2}+P_2\frac{di_1}{dt}+P_3i_1=0 \tag{37}$$

in which

$$\left.\begin{aligned} P_1&=\frac{R_1}{L_1}+\frac{1}{R_2C_b}+\frac{1}{R_pC_b'}\\ P_2&=\frac{1}{L_1C_a}+\frac{R_1}{R_2L_1C_b}+\frac{R_1}{R_pL_1C_b'}\\ P_3&=\frac{1}{R_2L_1C_aC_b}-\frac{1}{R_2L_1C_m^2}+\frac{1}{R_p}\left(\frac{1}{L_1C_aC_b'}-\frac{1}{L_1C_mC_m'}\right)\end{aligned}\right\} \tag{38}$$

Introducing the uncoupled damping factors and frequency of (36), equations (38) become

$$P_1=2\alpha_a+\frac{1}{R_2C_b}+\frac{1}{R_pC_b'}$$

$$P_2=\beta_a^2+\frac{2\alpha_a}{R_2C_b}+\frac{2\alpha_a}{R_pC_b'} \tag{39}$$

$$P_3=\frac{\beta_a^2}{R_2C_b}-\frac{1}{R_2L_1C_m^2}+\frac{1}{R_p}\left(\frac{\beta_a^2}{C_b'}-\frac{1}{L_1C_mC_m'}\right)$$

This circuit is singly periodic and its frequency is given by (12) when the condition for oscillation, (13), has been satisfied.

CRYSTAL BETWEEN GRID AND FILAMENT

IV. Tuned Plate Circuit. The circuit for this case is shown in Fig. 4 and the simplified diagram in Fig. 5, where C_2 is again $C^1_2+C_f$. C_p is here the coupling capacity instead of C_g. Writing the Kirchhoff relations, we have

$$L_1\frac{di_1}{dt}+R_1i_1+\frac{1}{C_3}\int i_1dt+\frac{1}{C_g}\int(i_1-i)dt=0 \tag{40}$$

$$L_2\frac{di_2}{dt}+R_2i_2+\frac{1}{C_2}\int(i_2+i_p-i)dt=0 \tag{41}$$

$$\frac{1}{C_p}\int idt-\frac{1}{C_2}\int(i_2+i_p-i)dt-\frac{1}{C_g}\int(i_1-i)dt=0 \tag{42}$$

$$i_p=\frac{1}{R_p}(e_p+\mu e_g) \tag{43}$$

where

$$e_p=-\frac{1}{C_2}\int(i_2+i_p-i)dt$$

$$e_g=\frac{1}{C_g}\int(i_1-i)dt$$

These equations differ only in minor details from (20) to (23), and the method of elimination is strictly analogous and will not be repeated. When the process has been carried out, there again results the fourth order differential equation

$$\frac{d^4i_1}{dt^4}+P_1\frac{d^3i_1}{dt^3}+P_2\frac{d^2i_1}{dt^2}+P_3\frac{di_1}{dt}+P_4i_1=0 \tag{44}$$

in which

$$\left.\begin{aligned}
P_1&=\frac{R_1}{L_1}+\frac{R_2}{L_2}+\frac{1}{R_pC_b''}\\
P_2&=\frac{1}{L_1C_a}+\frac{R_1R_2}{L_1L_2}+\frac{1}{L_2C_b}+\frac{1}{R_p}\left(\frac{R_1}{L_1}+\frac{R_2}{L_2}\right)\frac{1}{C_b''}\\
P_3&=\frac{R_2}{L_1L_2C_a}+\frac{R_1}{L_1L_2C_b}+\frac{1}{R_p}\left(\frac{1}{L_1C_aC_b''}+\frac{R_1R_2}{L_1L_2C_b''}-\frac{1}{L_1C_mC_m''}\right)\\
P_4&=\frac{1}{L_1L_2C_aC_b}-\frac{1}{L_1L_2C_m^2}+\frac{R_2}{R_p}\left(\frac{1}{L_1L_2C_aC_b''}-\frac{1}{L_1L_2C_mC_m''}\right)
\end{aligned}\right\} \tag{45}$$

Substituting uncoupled frequencies, damping factors and coupling coefficients, they become

$$\begin{aligned}
P_1&=2(\alpha_a+\alpha_b)+\frac{1}{R_pC_b''}\\
P_2&=\beta_a^2+4\alpha_a\alpha_b+\beta_b^2+\frac{1}{R_p}(\alpha_a+\alpha_b)\frac{2}{C_b''}\\
P_3&=2(\alpha_b\beta_a^2+\alpha_a\beta_b^2)+\frac{1}{R_p}\left[(\beta_a^2+4\alpha_a\alpha_b)\ \frac{1}{C_b''}-\frac{1}{L_1C_mC_m''}\right]\\
P_4&=\beta_a^2\beta_b^2\left[1-\tau^2+\frac{R_2}{R_p}\left(\frac{C_b}{C_b''}-\frac{C_m}{C_m''}\tau^2\right)\right]
\end{aligned} \tag{46}$$

where

$$\frac{1}{C_m}=\frac{C_0}{C_2C_g};\quad \frac{1}{C_m''}=\frac{1}{C_m}+\frac{\mu}{C_d};\quad \frac{1}{C_b''}=\frac{1}{C_b}+\frac{\mu}{C_m}$$

C_d is the capacitance between the points *dd* of Fig. 5. The other quantities have the same meaning as in Case I.

V. Inductance-Loaded Circuit. As in Case II, the coefficients $P_1 \cdots P_4$ are obtained by putting $C_2=C_f$. The circuit is doubly periodic.

VI. Resistance-Loaded Circuit. As in Case III, multiply (44) by L_2, put $L_2=0$, and there results

$$\frac{d^3i_1}{dt^3}+P_1\frac{d^2i_1}{dt^2}+P_2\frac{di_1}{dt}+P_3i_1=0 \tag{47}$$

where

$$\left.\begin{aligned} P_1&=2\alpha_a+\frac{1}{R_2C_b}+\frac{1}{R_2C_b''},\\ P_2&=\beta_a{}^2+\frac{2\alpha_a}{R_2C_b}+\frac{2\alpha_a}{R_pC_b''},\\ P_3&=\frac{\beta_a{}^2}{R_2C_b}-\frac{1}{R_2L_1C_m{}^2}+\frac{1}{R_p}\left(\frac{\beta_a{}^2}{C_b''}-\frac{1}{L_1C_mC_m''}\right) \end{aligned}\right\} \tag{48}$$

NUMERICAL CALCULATIONS

As pointed out above in (18), (19), (12), and (13), the expression for the frequency of the vacuum-tube-driven circuit, when doubly periodic, is given by

$$\beta^2=\frac{P_2\pm\sqrt{P_2{}^2-4P_4}}{2} \tag{49}$$

where the coefficients of the fourth order differential equation have such values that the condition for oscillation, i.e.,

$$\frac{P_2\pm\sqrt{P_2{}^2-4P_4}}{2}=\frac{P_3}{P_1} \tag{50}$$

is satisfied, and for the singly periodic circuit

$$\beta^2=P_2 \tag{51}$$

where

$$P_2=\frac{P_3}{P_1} \tag{52}$$

In satisfying the condition for oscillation for any given values of the uncoupled frequencies β_a and β_b, R_p is a variable parameter. R_p is, in reality, the reciprocal of the slope of the static characteristic of the tube as given by (23). Since, however, the actual characteristic of the tube is not linear but is curved, R_p has different values for various points on the curve. Moreover, for cyclic variations in grid potential, R_p has different values throughout the cycle and it must, therefore, when used in numerical calculations, be thought of as a sort of average value taken over the cycle. The greater the limits of grid potential variations, the greater is R_p. It is thus a purely fictitious quantity, to which it is impossible to give a definite physical meaning, but which serves a useful purpose. In general, the larger the value reached by R_p, the more vigorously is the circuit oscillating.

The direct method for computing the frequency of the system would be to substitute in (50) values of P's derived from (36) or (46), using any pair of values for the uncoupled frequencies β_a and β_b, and then to solve for R_p. This value of R_p, when substituted in (49), would then give the frequency of the driven system for these values of β_a and β_b. Unfortunately, (50) is too complicated a function of R_p to make this procedure possible, and a graphical method was employed. The right- and left-hand members of (50) were plotted as a function of R_p for a series of values of β_a and β_b. The intersection of these curves then gave the value of β^2 for this combination of uncoupled frequencies.

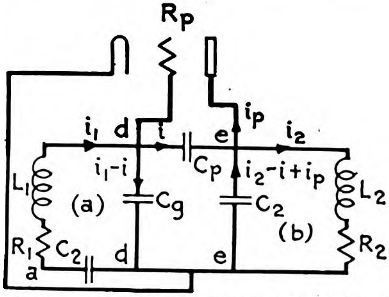

Fig. 5—Simplified Diagram for Fig. 4.

To simplify the computations and make them of more general application, relative frequencies rather than actual frequencies of any particular crystal were used. Dividing (49) and (50) by β_a^2 we have

$$\left(\frac{\beta}{\beta_a}\right)^2 = \frac{\dfrac{P_2}{\beta_a^2} \pm \sqrt{\left(\dfrac{P_2}{\beta_a^2}\right)^2 - \dfrac{4P_4}{\beta_a^4}}}{2} \qquad (53)$$

Since P_2 and P_4 from (36) are functions of β_a and β_b, (53) gives the ratio of the driven frequency of any crystal to its undriven value in terms of the ratio of the uncoupled frequency of the plate circuit to the undriven frequency of the crystal, the undriven frequency of the crystal, damping factors and other constants of the circuit. In carrying out the computations β_a was taken as unity.

The computations were carried out for a crystal having the dimensions shown in Table I. The measured value of its driven fundamental frequency for "thickness" oscillations was 451.53 kc.

TABLE I

TUNED PLATE CIRCUIT

Dimensions of crystal	Electrical Constants of crystal	Tube constants (UX–201A)
length = 3.328 cm.	$L_1 = 3.656$ h	G-F cap. = 5.8 $\mu\mu$f
breadth = 2.750 "	$R_1 = 9035.5$ Ω	G-P " = 10.1 "
thickness = 0.6361 "	$C_1 = 5.7551$ $\mu\mu$f	P-F " = 6.1 "
	$C_2 = 0.03165$ "	Amp. fact. = 8.5 "

Coupling coefficients:

Crystal between grid and plate, $\tau^2 = 1.0356 \times 10^{-6}$

Crystal between grid and filament, $\tau^2 = 3.1355 \times 10^{-6}$

$\frac{\beta_b}{\beta_a}$	R_p	$\frac{\beta}{\beta_a}$ Crystal between grid and plate	$\frac{\beta}{\beta_a}$ Crystal between grid and filament	$\frac{\beta}{\beta_a}$ Natural frequencies undamped
0.94	47000	1.0000121	—	1.0000035
0.96	72500	1.0000122	—	1.0000060
0.98	144000	1.0000143	—	1.0000120
0.99	260000	1.0000262	—	1.0000251
0.995	370000	1.0000540	—	1.0000523
1.005	455000	—	0.999847	0.999846
1.01	394000	—	0.999922	0.999921
1.02	238000	—	0.999963	0.999959
1.04	126000	—	0.999986	0.999979
1.06	87000	—	0.999996	0.999986

In this table, the column marked R_p gives the effective resistance of the tube for stable oscillations, that is, the value for which one of the damping factors of the driven coupled circuit

AUTHOR'S NOTE:—In computing the equivalent electrical constants for the crystal the following unpublished formulas were used which were reported by Cady to the executive committee of the American section U. R. S. I. in 1926.

$$L_l = 130\,\frac{le}{b} \qquad C_1 = 0.40\,\frac{bl}{e} \qquad R_1 = 130{,}000\,\frac{e}{bl}$$

$$L_e = 130\,\frac{e^3}{bl} \qquad C_2 = 0.0022\,\frac{bl}{e}$$

Where l, b, e represent respectively the length, breadth, and thickness of the quartz plate, measured in directions which are respectively normal to the electric and optic axes, along the optic axis and along the electric axis. L_l and L_e are the equivalent inductances for lengthwise and thickness oscillations, respectively. The computations, the results of which are given in Table I, were well under way when Van Dyke's paper, loc. cit., appeared, and since the values given by the newer formulas differ from the above by only a few per cent, it was deemed not worth while to repeat the computations, as they are extremely laborious.

out these computations, it is found that with the above circuit constants for ratios of β_b/β_a less than unity, the condition for oscillation can be satisfied only when the plus sign of (53) is used, and for ratios greater than unity, with the minus sign only.

These results are shown in Fig. 6, in which the dash-dot curve shows the frequency of the vacuum-tube-driven crystal-controlled oscillator as a function of the relative frequency of the plate circuit, while the dot-dot curve shows the natural frequencies of the same system if it were without resistance. It is seen that the effect of driving is to increase somewhat the frequency of the system for both connections of the crystal, and that this effect is greater the farther off resonance the circuits are. The experimentally-measured frequencies are shown in the full-line curve. It was found impossible to satisfy the condition for oscillation using the value of R_1 obtained by Van Dyke's formula, even though R_2 were as small as 0.3 ohm, the value used in the experimental work. A series of calculations was carried through, using successively smaller and smaller fractions of R_1, and in the above curves R_1 was put equal to one-tenth that given by Van Dyke's formula. It will be noted that the experimentally-measured curves show a somewhat sharper resonance effect than the computed ones, and if a still smaller value of R_1 had been used the check would have been better. Because of the uncertainty in the values of some of the other constants of the circuit, this can hardly be considered a method for measuring the effective resistance of a crystal, but it indicates that the values given by Van Dyke's formula are probably too large.

INDUCTANCE-LOADED CIRCUIT

Calculations for this case are not given since, as pointed out above, this is a special case of the tuned plate circuit in which C_2 is simply the capacitance between plate and filament of the tube.

RESISTANCE-LOADED CIRCUIT

Introducing relative frequencies in (51) and (52), we have

$$\left(\frac{\beta}{\beta_a}\right)^2=\frac{P_2}{\beta_a^2}=\frac{\dfrac{P_3}{\beta_a^2}}{P_1} \qquad (53)$$

Substituting the values for $P_1 \ldots\ldots P_3$ of (38) and (48) in (53), it was found that the condition for oscillation can be satisfied only for the case of the crystal between grid and plate, and for this the following results were computed for different values of R_2 the load, or series plate resistance.

TABLE II

R_2	$\frac{\beta}{\beta_a}$	R_p
10000	1.00000536	6240
20000	1.00000503	6000
30000	1.00000096	5900
50000	1.00000048	5750
100000	1.00000042	5700

In this case it is to be noted that the crystal is forced from its natural frequency, at most, by only five parts in a million, and that the variation in frequency is very much less than in the case of the tuned circuit load. The values of R_p, however, are relatively small, indicating that the oscillations are weak, and the fact that they decrease as R_2 is increased indicates that the oscillations become weaker as the load is increased.

Experimental Checks

To test the theory as given above, two crystal-controlled oscillators of the various types there discussed were set up, each having its own *A* and *B* battery supply, and appropriate precautions were taken to insure a minimum of interaction between them. Each was loosely coupled to a third circuit containing a rectifier and an audio-frequency amplifier. Connected to this third circuit was an audio oscillator of continuously variable frequency calibrated in terms of a series of standard tuning forks. The crystals under test were ground in pairs so as to give audio-frequency beats of 300 to 500 cycles between them. One crystal-controlled oscillator was used to maintain a constant radio frequency while the other was experimented upon. The beat frequency between the crystals was then matched by the audio oscillator. This double heterodyne arrangement is an exceedingly sensitive method for studying problems involving frequency changes, since a variation of one cycle per second in either of the radio-frequency circuits gives an audio-frequency beat. The circuits under test were sufficiently stable to permit the audio-frequency beats to be held constant to one beat in 10 seconds.

The results for the tuned plate circuit case are shown by the full-line curve of Fig. 6, reference to which has already been made. For the resistance-loaded circuit the variations of frequency with load resistance were so small as to be barely within the limit of error of measurement, i.e., one or two parts per million. This checks the results of Table II, where maximum changes of only 5 parts per million were predicted. For standardization purposes, where extreme accuracy is the objective, the resistance-loaded circuit is thus much to be preferred. However, this circuit gives much weaker oscillations and will in general require an extra stage of amplification. Moreover, a crystal must be a very "good oscillator" to be used at all in this circuit. Out of some twenty crystals used in connection with this work, only three would function with this arrangement.

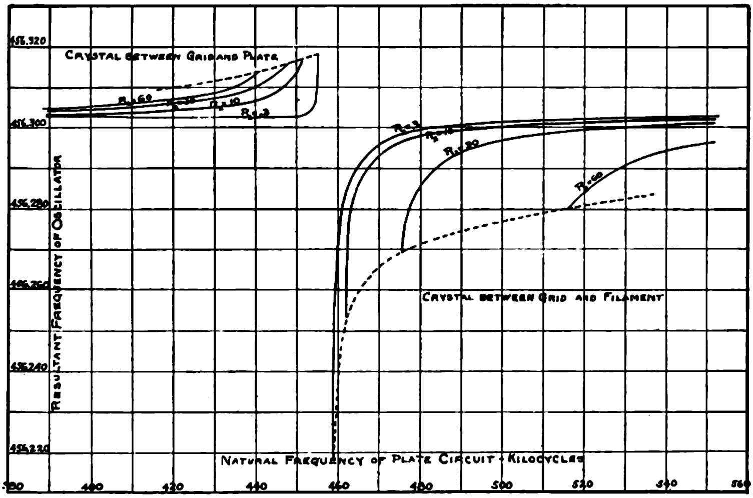

Fig. 7—Variation of Driven Frequency with Resistance of Tuned Plate Circuit.

An experimental study was made of the effect of varying the resistance R_2 for the two cases of tuned plate circuit oscillators, and the results are shown in Fig. 7, where the frequency of the driven system is shown in terms of the uncoupled frequency of the plate circuit. These results indicate that as R_2 is increased the separation between coupled frequencies is increased and that the sharpness of resonance is decreased. The range of plate circuit frequency over which oscillations occur is decreased by increasing load.

CONCLUSIONS

A quartz-crystal oscillator, when used to stabilize a vacuum-tube-driven circuit, does not oscillate at a frequency determined by its elastic and piezo-electric properties alone, but becomes part of a coupled system, and the actual resultant frequency is influenced by the degree of coupling of the two systems and the values of the constants of the entire circuit, including those of the driving device in the case of continuous oscillations. In doubly periodic vacuum-tube-driven circuits, one of the normal modes of oscillation is excited when the crystal is connected between grid and plate, and the other when connected between grid and filament. Although the oscillations are more powerful when the frequency of the plate circuit is close to that of the crystal, the departures of the resultant frequency from the natural frequency of the crystal are greater. For purposes of accurate frequency standard maintenance, the resistance-loaded circuit is much to be preferred, and when a crystal has been standardized it must always be used in exactly the same circuit and under exactly the same conditions as when the standardization was made. It is desirable from this standpoint to preserve not merely the crystal, but the entire circuit permanently assembled.

In conclusion, the author wishes to express his indebtedness to the American Association for the Advancement of Science for a grant for the purchase of equipment, to Dr. A. Hoyt Taylor of the Naval Research Laboratory for a supply of accurately-ground quartz plates, and to Mr. J. C. Cavender for carrying out the experimental measurements.

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QUANTITATIVE METHODS USED IN TESTS OF BROADCAST RECEIVING SETS*

By

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Summary—The general classes of receiving set measurement used by the authors are outlined as special engineering tests and production tests. The apparatus and methods for both are described. A new form of radio-frequency oscillator, designed for this work. is described. Shielded test booths used for receiving-set measurement work are described. The importance of care in selecting the conditions of test used in making quantitative measurements on a receiving set is emphasized.

MOST electrical quantities encountered in general electrical engineering are readily measurable, so that its practice has long been exact in measurements and tests of electrical products. Dynamo-electric machinery and power transformers, for example, have long been studied with the aid of relatively simple meter equipment. Radio engineering practice has not been so fortunate, except in transmitter equipment where quantities are similar to those of general electrical practice. In radio receiving set engineering, numerous obstacles to quantitative measurement work were present which required the development of new methods and new apparatus. These problems were caused in the main by the fact that the electrical quantities involved were so small that previously used meters and methods were useless with them. Many of the voltages and currents to be measured in receiving set practice are but millionth parts of those met in general electrical practice.

As a result of this lack of means of measurement, receiving set tests during the first twenty-five years or so of the radio art were conducted in a necessarily crude, practical way, chiefly by so-called "listening tests," wherein the receiving set was operated exactly as in actual service, and observations made by ear. This usually required a comparison test which rated the receiver under test in terms of some other receiver, the performance of which was familiar through previous experience. Most of the resulting data was far from having engineering

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exactitude. Signals delivered were said to be "one hundred times audibility," a "half-stage audio better" than the standard receiving set, "audible two rooms away," etc. Selectivity was "sharp as a knife," or even "razor-edge."

Using such methods of test and criteria, it was necessary to operate the receiving set in several locations having different signal conditions, and to spend considerable time at each location, in order to secure even approximate information on the performance characteristics of the receiver. Results at best were inexact and had little possibility of usefulness in predicting the performance of the receiving set under other conditions.

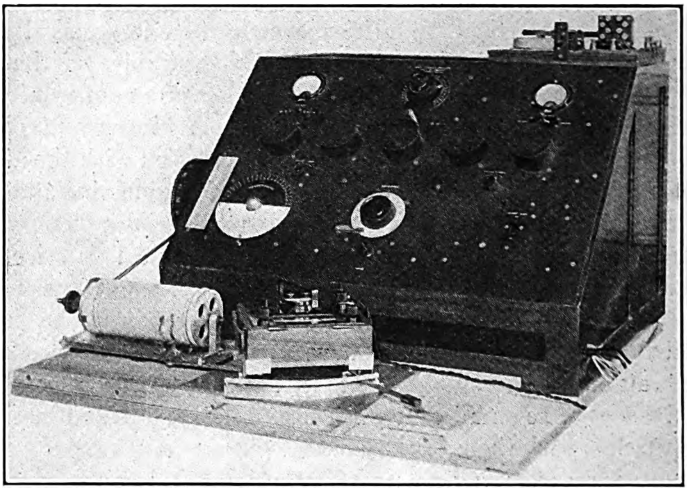

Fig. 1—Audio-Frequency Oscillator.

The ideal method of measurement would furnish such data as will make it possible to predict the performance of the receiving set under any specified service conditions. Such methods are now available for the measurement of sensitivity, selectivity, and fidelity of receiving sets, and it is the object of this paper to describe some of these methods. With these methods, a measurement laboratory is able to apply to a receiving set a measured input signal, and to measure the receiver output. In effect, the laboratory is supplied with a transmitting station which will transmit when desired, on any required carrier frequency, with any desired power, and with music or single tone modulation of fidelity equal to that of the best broadcast stations, In the year 1922, the authors began work to develop methods and equipment for quantitative measurement of receiving set performance. At that time considerable was known about audio-amplifier measurements, and some work had been done by various laboratories and individuals in the measurement of radio-frequency amplifiers, but no effective method for measuring overall performance of receiving sets had been developed. In 1924, the work was greatly furthered by the assistance of Dr. Walter Van B. Roberts, and one system developed by him is described later in this paper.

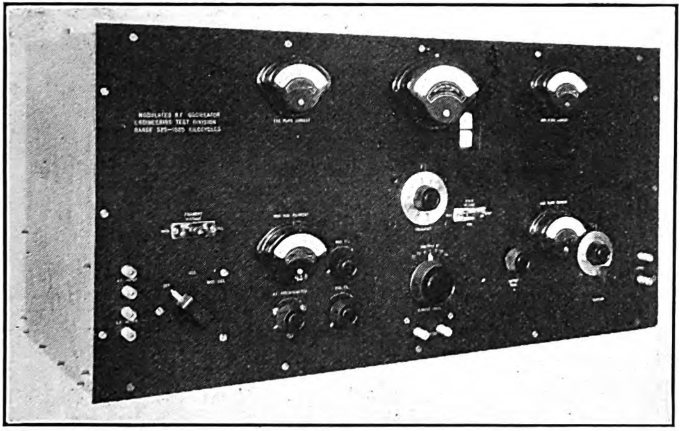

Fig. 2—Engineering Test Radio-Frequency Oscillator.

Early in this work it was found, as would be expected, that the determination of receiving set performance by measurement of individual parts of the set (for example, the radio and audio amplifiers) was impracticable. This was chiefly due to the fact that results are usually different when reactions between parts are absent, as they are under measurement of parts individually. Additional reasons arise from the facts that the electrical constants of most receiving sets are changed seriously when connections are made to internal parts, and that more time is required for measuring parts than is needed for a single over-all test. Sometimes, measurement of individual parts is desirable or even necessary, however, as when attempting to locate the cause of inferior performance of a set, or when engaged in development and design of a new set, in which cases segregation of the effects of the several parts is necessary. The methods of measurement to be described are intended chiefly for determination of the effectiveness of a complete receiving set, and are so arranged that the test results can be used to predict how the set will perform under any specified service conditions.

In order to obtain useful and dependable measurement results, several important conditions must be met, namely:

1. The equipment must be built so as to be accurate, reliable, and easy to operate.

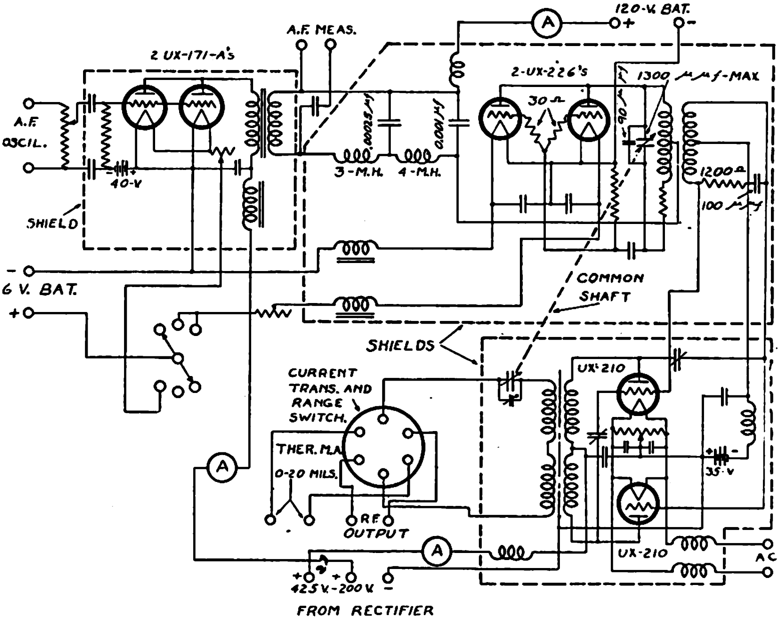

Fig. 3—Circuit of Engineering Test R. F. Oscillator.

2. Test conditions must represent service reception conditions, and no conditions can be present which cause erroneous results or conclusions.
3. The method of test should be suited to the use to which the results are to be put.

The authors have found that two types of measurement equipment are required to meet radio engineering needs. One is that used in engineering laboratories, in connection with development, design, or investigation of receiving set problems. The other is that used as a final factory production test, and applied to large numbers of sets. In the latter equipment, accuracy can be sacrificed to some extent, but in its place must be substituted greater simplicity of adjustment and operation, and increased ruggedness of construction. Reliability must be had in both classes of equipment to a maximum degree. On account of the difference between the equipments used for the two purposes, they will be discussed separately, and the test conditions which are of importance in connection with each will be discussed following the description of the apparatus in each case.

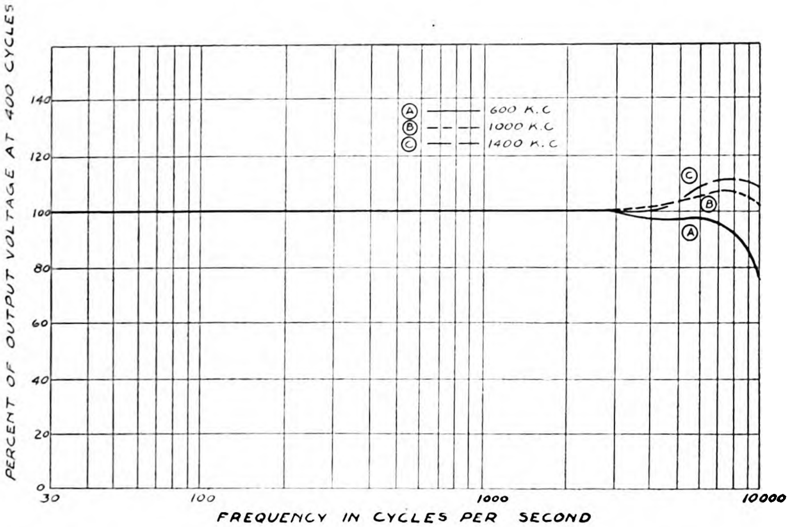

Fig. 4—Overall A. F. Characteristics of Engineering Test R. F. Oscillator.

Engineering Test Equipment

Audio-Frequency Oscillator. The source of audio-frequency voltage which we have found most satisfactory is a beat frequency vacuum-tube oscillator, employing the difference frequency between two radio-frequency oscillators. Such an oscillator is shown in Fig. 1. By combining and rectifying the radio frequencies, and then amplifying the resultant difference frequency, it is possible to obtain from the oscillator shown in this figure audio-frequency voltages of the order of 50 volts. The wave shape of the oscillations does not contain more than 1 or 2 per cent of harmonics even at the lowest frequencies, and the output potential will remain effectively constant over the frequency range of approximately 30 to 10,000 cycles per second, without readjustment.

The output voltage from this oscillator is sufficient to apply the necessary potential directly to the grids of the modulator tubes of the r.f. oscillator, without requiring any intervening amplification. In front of the oscillator in Fig. 1 will be seen the automatic curve drawing mechanism used for taking fidelity characteristics. The construction and use of this mechanism has already been described in these PROCEEDINGS.[1]

Radio-Frequency Oscillator. Two radio-frequency oscillators have been used in our engineering test work. One has already been described in a paper by Messrs. Rodwin and Smith of this laboratory. This oscillator is an excellent device and is capable of wide usefulness in test work. Those desiring to obtain the details of its design are referred to the above paper,[2] which also contains information on methods of oscillator calibration.

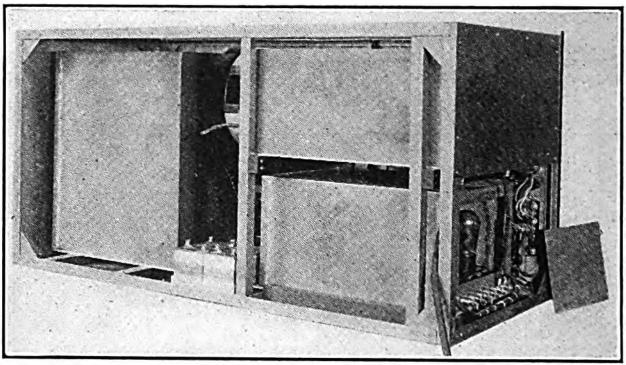

Fig. 5—Back of R. F. Oscillator Showing Separate Shielded Compartments.

Another r.f. oscillator has been recently completed, and is shown in Fig. 2. This oscillator embodies certain changes in construction as well as some new design features, which give improved operation. It uses two UX-171-A tubes connected in parallel as modulators, two UX-226 tubes in parallel as oscillators, and two UX-210 tubes connected in push-pull arrangement as r.f amplifiers. The modulator and oscillator tube filament currents are supplied from a storage battery. The amplifier tube filaments are supplied with alternating current at the proper voltage. The modulator and amplifier tubes receive their plate supply from a heavy duty "eliminator" operated from a.c., with its output carefully filtered. The oscillator tubes receive their plate supply from a bank of storage cells.

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[1] E. T. Dickey, "Notes on the Testing of Audio-Frequency Amplifiers," PROC. I.R.E., 15, 687; Aug. 1927.

[2] George Rodwin and Theodore A. Smith, "Radio-Frequency Oscillator for Receiver Investigations," PROC. I.R.E., 16, 155; Feb. 1928.

The r.f. range of the oscillator is 525 to 1525 kc. Slightly more than the broadcast range is provided for convenience.

It is desirable that the oscillator frequency shall not be appreciably affected by changes in the power supplied to the output circuit. By careful neutralization of the r.f. amplifier circuit, the effect of variation (from maximum to minimum output) on the frequency of the oscillator is reduced to approximately 0.01 per cent at the high-frequency end of the range, and is lower than this at the opposite end. This amounts to no more than 150 cycles variation.

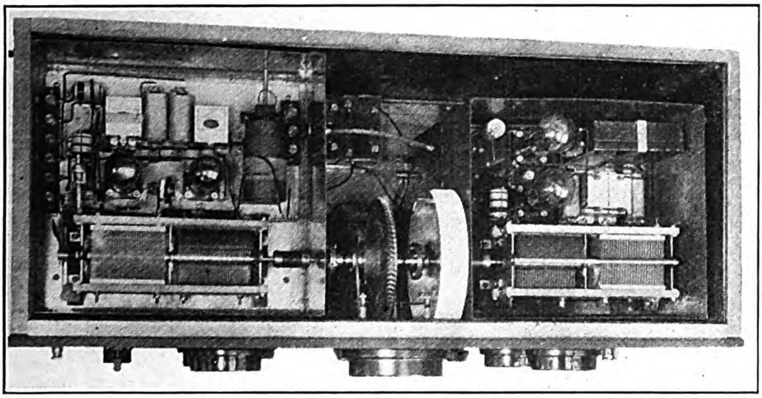

Fig. 6—Top of R. F. Oscillator Showing Oscillator and Amplifier Compartments with Shield Covers Removed.

The circuit for the oscillator is shown in its essential features in Fig. 3. It will be noted that grid resistors are used in series with the grids of the oscillator tubes. These are necessary to avoid parasitic oscillations. The size of the oscillator grid condenser is quite critical if the best audio-frequency characteristic is to be obtained. In order to prevent excessive dropping of the audio-frequency characteristic at the high-frequency end of the audio

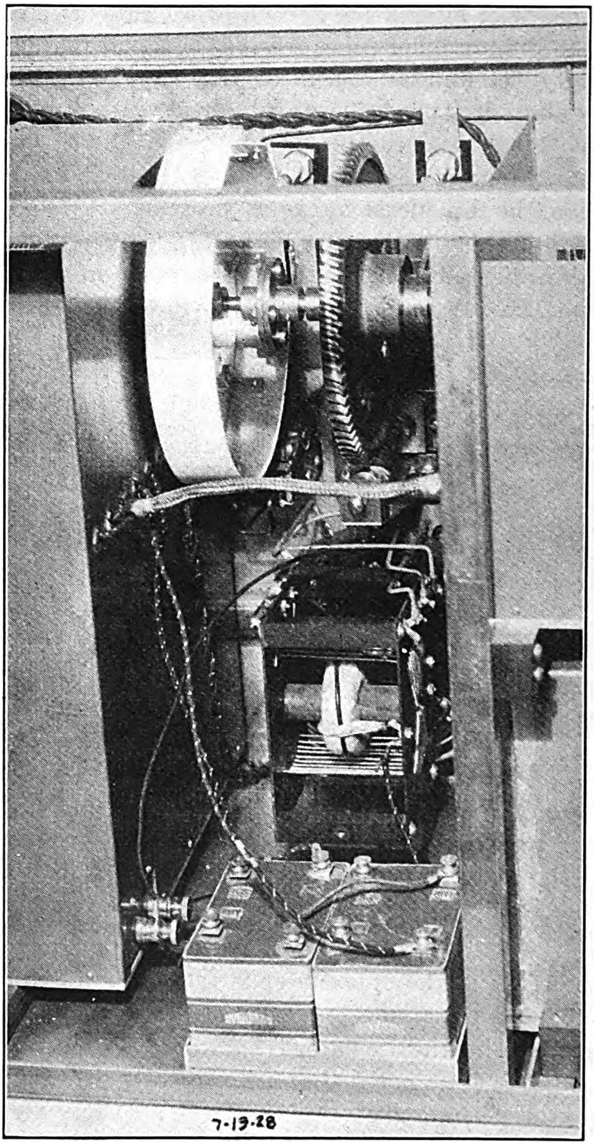

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Fig. 7—View of Condenser Worm Drive and Current Transformer.

range, it is necessary to insert resistance in series with the oscillator circuit. The characteristics of the oscillator with regard to the audio-frequency modulation of the r.f. output and taken with constant input a.f. voltage, are shown in Fig. 4, for three frequencies in the broadcast range. It will be seen that a compromise has been reached at the high audio frequencies between the characteristic at 600 kc and that at 1400 kc. When the a.f. characteristics of most receiving sets are considered, it is evident that these variations in the r.f. oscillator characteristic can be neglected in practically all cases.

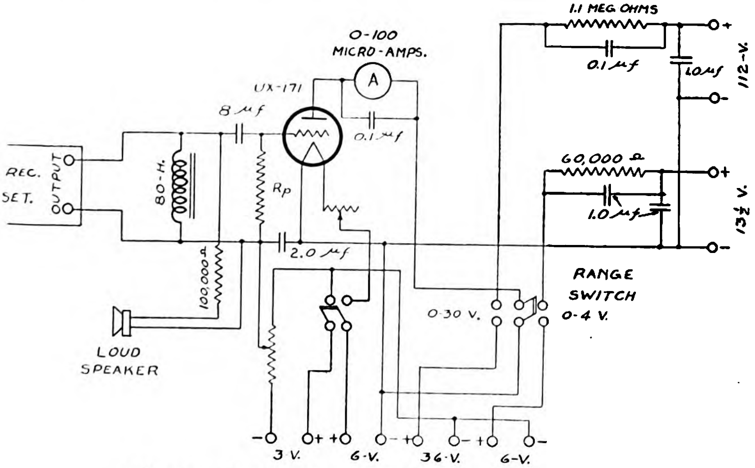

Fig. 8—Engineering Test Output Meter Circuit.

Modulation is obtained by introducing audio-frequency voltage in series with the plate-supply voltage of the oscillator tubes. This arrangement has been found to give quite satisfactory results, and it appears to have advantages over the constant current system for test oscillator use. The same a.f. voltage will produce equal percentage modulation at all radio frequencies. By properly connecting the primary and secondary of the modulation transformer, it is possible to make the plate current of the modulator tubes counteract the magnetizing effect of the oscillator tube plate current, thus avoiding any tendency toward core saturation due to direct current. This permits the use of a transformer having very low leakage reactance which is of considerable aid in obtaining a good audio-frequency characteristic for the modulation system.

Separate shielded compartments for the component parts, together with a complete shield surrounding the entire oscillator, have been provided as shown in Figs. 5 and 6. The compartment shields have been left "floating," i.e., ungrounded. This double shielding, together with the use of astatic wound coils in parts of the circuit carrying large r.f. currents, reduces the direct radiation from the oscillator to a minimum. The apparatus enclosed in each shielded compartment is shown roughly by the dotted lines in Fig. 3. It will be noted that the filter system in the secondary of the modulation transformer is inside of the oscillator compartment shield. This effectively prevents the possibility of radio-frequency current getting back into the modulator system, or out to the modulation voltage measuring meter.

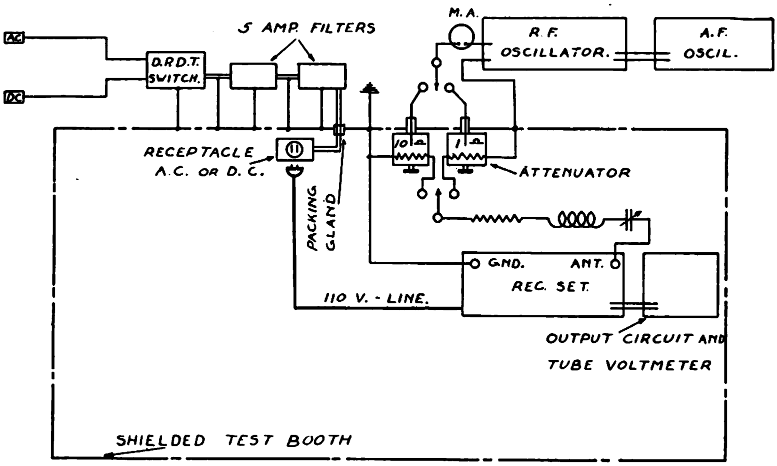

Fig. 9—Apparatus Arrangement in Test Booth.

Note: All leads between filters and switch and line boxes are carefully shielded and shielding connected to ground.

The coupling between the oscillator inductor and the inductor feeding the r.f. amplifier tubes is fixed. The resistor and condenser shunted across the secondary winding have their values so chosen that the amplitude of the radio-frequency oscillations is maintained practically constant throughout the broadcast range. The values of resistance and capacitance used are, of course, dependent on the coupling between the primary and secondary windings.

The plate-circuit coils of the r.f. amplifier system are wound astatically, and variable coupling is obtained between them and the output circuit by means of a second pair of movable astatic coils. An electro-static shield is provided between the two sets of coils to permit zero coupling to be obtained. The output circuit is tuned to the oscillator frequency in order to reduce harmonics in the output current. The use of two r.f. amplifier tubes connected in push-pull also assists in this reduction. The output and meter resistance make this circuit sufficiently broad so that it does not have any material effect in cutting side bands.

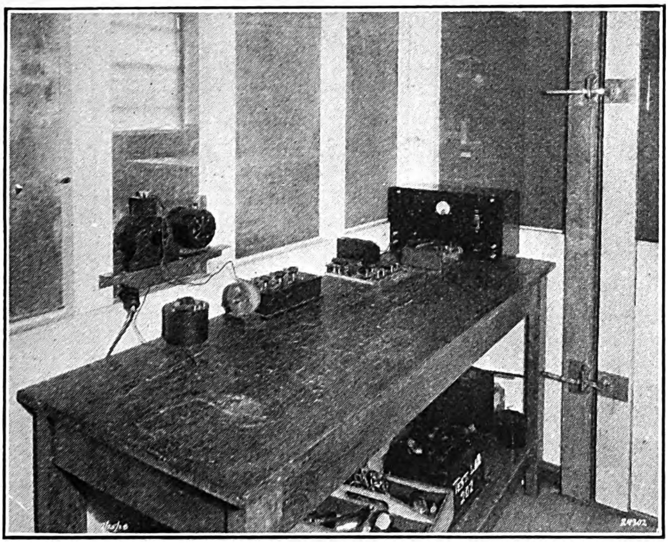

Fig. 10—Interior of Engineering Test Booth.

The oscillator variable condenser and the main variable condenser of the output circuit are directly connected to the same shaft. The large worm gear drive used for these condensers is shown in Figs. 6 and 7. This is especially designed to avoid back lash, and permits very accurate frequency calibration of the frequency adjustment dial. The main tuning variable condensers are of the straight line frequency type. This permits the vernier dial, which is attached to the worm drive, to be calibrated in divisions representing a small number of kilocycles, for use in taking selectivity characteristics. With the worm gear and this vernier dial, it is possible to detune the oscillator in steps of 2 kc either side of a test frequency, with a maximum possible error with respect to the test frequency of not more than 300 cycles, at any point up to 100 kc, from the test frequency. The worm drive permits variation of frequency as far as desired, of course, but selectivity curves are generally not taken beyond 100 kc, either side of resonance.

A high accuracy of oscillator frequency is assured by the use of relatively low inductance and high capacitance in the oscillator circuit. This tends to minimize the effect on the frequency, of variations in filament and plate voltage, as well as the changing of tube characteristics over a long period of time.

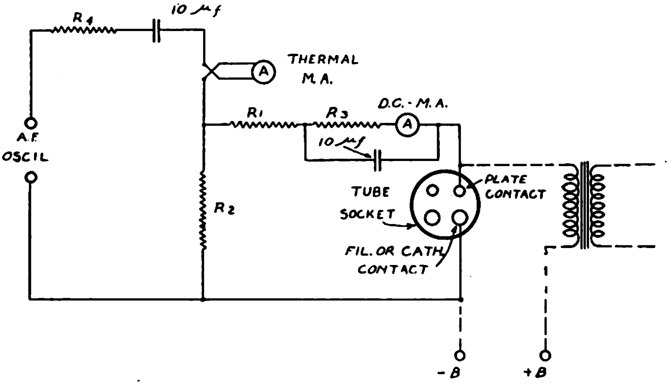

Fig. 11—Input Circuit for A. F. Amplifier Tests.

In Fig. 7 the current transformer will be seen below the condenser gear. This transformer is similar in general to that described in the previously mentioned paper by Messrs. Rodwin and Smith. The use of an electro-static shield between the primary and secondary windings, together with a "figure 8" winding for the secondary, has resulted in producing a transformer having an accuracy of not less than 97 per cent at any point of the broadcast range. This transformer is designed with an elaborate switching arrangement which provides six ranges from 50:1 to 1:50. With this transformer, a 20-milliampere thermal meter can be used to measure accurately output currents as low as 80 microamperes, and as high as the full output of the oscillator. The full output of the oscillator at 600 kc is approximately 80 milliamperes with 10-ohm output circuit, and at 1500 kc the maximum output is approximately 200 milliamperes.

Simplicity of operation has been retained to the greatest possible extent in this device to permit the taking of test data easily and quickly. Complete metering of all d.c. circuits is provided by meters on the front panel. The thermal meters for measuring the a.f. modulation voltage and the r.f. output current are connected externally, and are usually located on the table directly in front of the oscillator.

Transfer Circuit. For testing receiving sets designed for operation with antenna and ground connections, resistance type attenuators are used between the r.f. oscillator and the artificial antenna circuit. In order to cover accurately a wide range of voltages, two attenuators are used. One attenuator has an attenuation range of from 0.01 to 0.95 in seven steps, and an output resistance of 1 ohm. The second attenuator has a range of from 0.1 to 9.0 and an output resistance of 10 ohms. Using these attenuators in connection with the above described oscillator, it is possible to impress (in series with the artificial antenna circuit) voltages of from 0.8 microvolts up to 0.7 volts. This range is sufficient for all but very exceptional test requirements.

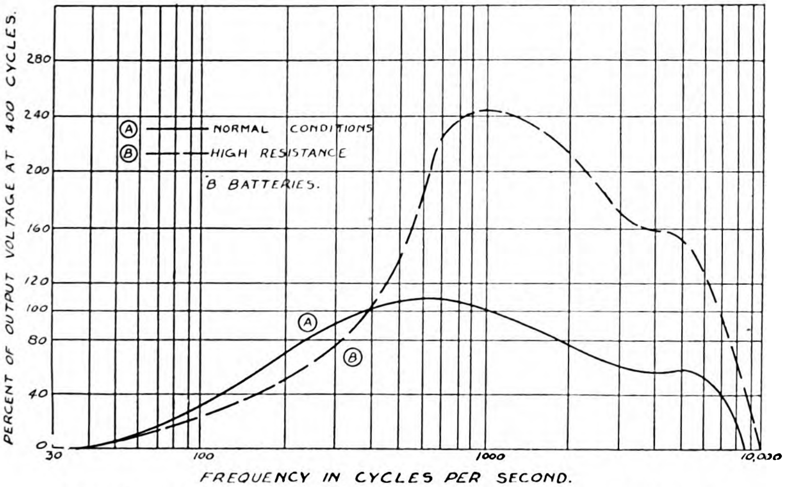

Fig. 12—Effect of High Resistance *B* Batteries on Fidelity of A. F. System.

For receiving sets arranged for loop operation, the output of the r.f. oscillator is connected to a small coil of known dimensions, and the loop of the receiving set is placed in the radio-frequency field of this coil. A single-turn coil approximately 2 inches in diameter is used for small field strengths, and a 4-turn coil approximately 8 inches in diameter for greater field strengths. The field strength at the receiving loop is then calculated from the dimensions of the transmitting coil, the current through it, and the distance between it and the receiving set loop.

Output Meter. A vacuum-tube voltmeter having two ranges (0–4 and 0–30 volts) is used to measure the voltage across the

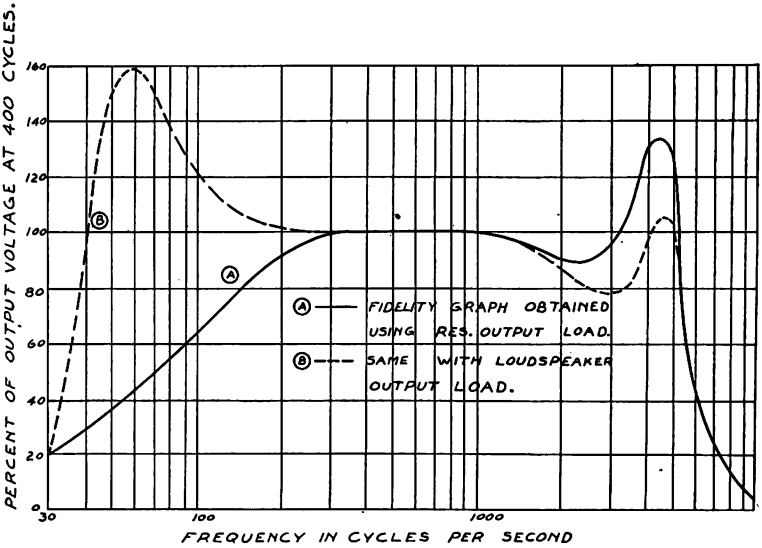

Fig. 13—Effect of Output Load on Fidelity Graph.

output resistance of the receiving set under test. Fig. 8 shows the circuit of the vacuum-tube voltmeter, together with the choke and condenser arrangement used in the receiving set output circuit when the latter has d.c. in its output circuit. The choke and condenser can, of course, be eliminated if there is no d.c. in the output circuit. The value of the resistor R_p is made equal to the plate resistance of the output tube. Across the output of the receiving set is connected a high impedance loudspeaker, in series with a 100,000-ohm resistor. This combination has no appreciable effect upon the output voltage, and serves to give a weak monitoring signal. This enables the tester to have an audible indication of proper performance during the test. It is very useful in checking undesired test conditions, such as oscillation in the receiving set, beat-note whistles between the r.f. oscillator and a powerful local broadcasting station, etc. If such a monitoring arrangement is not used, effects such as those mentioned above may be present and give erroneous test measurements, without the tester being aware of their presence.

The ranges provided by this tube voltmeter are sufficient for any of the usual measurements on receiving sets. If it is desired to measure voltages either above or below the useful range of this meter, a resistance divider or amplifier can be used at the input of the tube voltmeter to increase its range. This tube voltmeter has been found to be quite accurate and reliable, retaining its calibration within approximately 1 per cent over long periods of time. It is considerably more rugged than a thermo-couple instrument, since it is so designed that burnout of the plate meter is practically impossible. Compared against a thermocouple meter on a.c. voltages of bad wave shape, this tube voltmeter has an error no greater than plus or minus approximately 4 per cent, even with voltages having a total harmonic content of as much as 40 per cent. A zero reading of 10 microamperes is maintained in the plate meter by adjustment of the potentiometer. Once adjusted it will hold this zero setting for a considerable time. Those interested in obtaining further data on this tube voltmeter are referred to the previously mentioned article on a.f. amplifier testing.

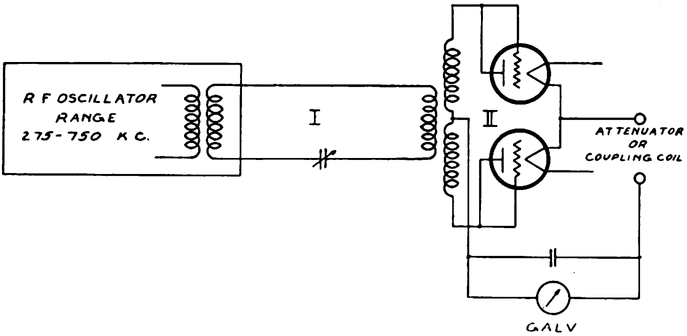

Fig. 14—Production Test R. F. Oscillator Theoretical Circuit.

ENGINEERING TEST CONDITIONS

Test Booth. It is important that the receiver under test be removed from the influence of extraneous effects such as power-line induction, atmospherics, stray pickup from local broadcasting stations, etc. The most effective way of accomplishing this is to place the receiving set on the inside of a well shielded booth. The r.f. and a.f. oscillators are situated on the outside of the booth. The general arrangement is shown diagrammatically in Fig. 9. The booth used for engineering tests at this laboratory has copper sheeting on the floor and parts of the sides. The top, front, and remainder of the sides are covered with copper screening. All the joints are lapped and well soldered. The copper screening was given a tinning coat before being erected to insure good electrical connection of the strands at each cross-over in the mesh. Metal weather stripping was used all around the door, and "refrigerator" type door clamps were used to secure sufficient pressure on the weather stripping to insure tight electrical contact when the door is closed.

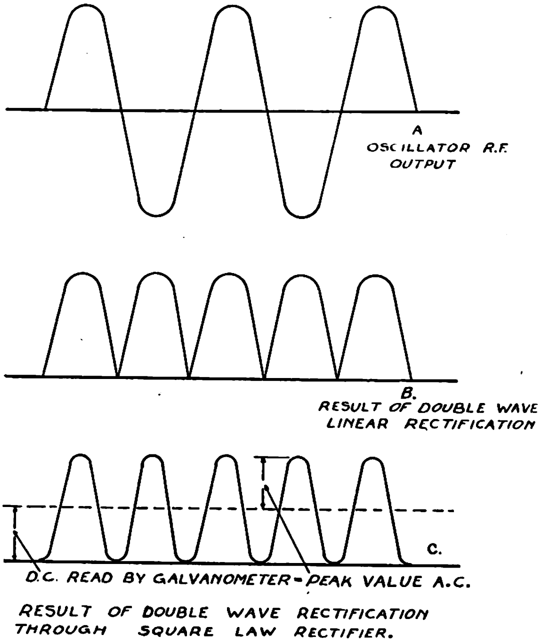
A
OSCILLATOR R.F. OUTPUT
B.
RESULT OF DOUBLE WAVE LINEAR RECTIFICATION
C.
D.C. READ BY GALVANOMETER
PEAK VALUE A.C.
RESULT OF DOUBLE WAVE RECTIFICATION THROUGH SQUARE LAW RECTIFIER.

Fig. 15—Production Test R. F. Oscillator Frequency Doubling Action.

Fig. 10 shows an interior view of one corner of the test booth. In this view at the right will be seen the door clamps, at the left the resistance attenuators and artificial antenna circuit, and at the end of the test table the output meter.

As will be seen from Fig. 9, the leads supplying alternating current to the inside of the booth are run through a system of filter circuits before entering the booth. The purpose of these is to prevent transmission of radio signals, line interference, etc., to the receiving set over these wires. Lighting lamps and their wires are kept outside the screen, so that induction from them may be avoided.

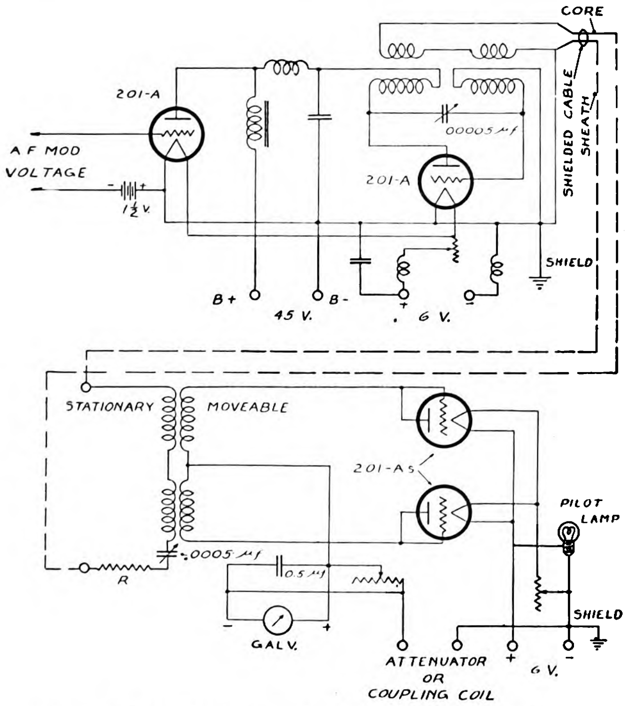

Fig. 16—Production Test R. F. Oscillator Circuit.

The position of the antenna, ground, and output circuit wires with respect to the receiving set is important. There is no general rule governing the position of these leads in actual use of the receiving set, of course, but it is best during test so to place these leads that they have a minimum effect upon the measurements. In general this involves avoidance of an arrangement which causes any of these leads to run close to the receiver for any considerable distance, and the simplest arrangement is generally to run the leads away from the receiving set in a direction perpendicular to the side from which they emerge.

Test Procedure. The test procedure used at this laboratory for taking measurements on receiving sets is the same as that outlined in the recently published Preliminary Draft of Report of the Committee on Standardization of the I.R.E. for 1928. Since this draft is available, no discussion on this subject will be given in this paper.

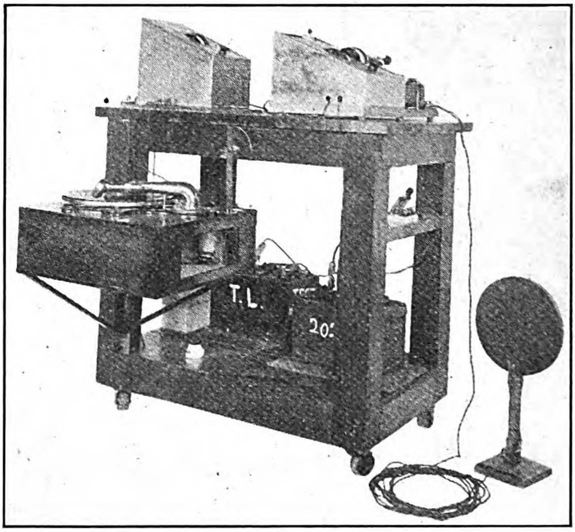

Fig. 17—Production Test R. F. Oscillator.

A. F. Amplifier Characteristics. It is generally desirable to take a fidelity characteristic of the a.f. amplifier system in a receiving set, to obtain information needed to analyze the overall fidelity characteristic in case features of this characteristic require study or correction. Segregation of the a.f. amplifier is usually sufficient. The general methods of test used at this laboratory have already been described in these PROCEEDINGS in the paper on a.f. amplifier testing mentioned previously. A circuit which has been found to be very convenient for plugging into a detector tube socket is shown in Fig. 11. This circuit is arranged to maintain the proper direct current through the primary of the a.f. transformer, using the normal *B* supply of the receiving set. At the same time it provides the proper resistance in series with this transformer, to simulate the plate resistance of the detector tube. Referring to Fig. 11, R_1+R_2 is made equal to the plate resistance of the detector tube. R_3 is adjusted to the proper value to give normal direct current through the transformer primary winding. R_3 and the d.c. meter are shunted with a 10-microfarad condenser so that they do not introduce appreciable impedance in the circuit for the a.f. voltage. R_2 is usually quite small with respect to R_1. R_4 is not absolutely essential, but where R_2 is small, it is desirable to have a somewhat higher resistance in the output circuit of the a.f. oscillator to prevent wave-form distortion. R_4 should be roughly equal to twice the plate resistance of the a.f. oscillator output tube.

General. In order to get complete data on a receiving set, it is necessary to test other functions which are not covered by the usual standard tests. A receiving set designed for power-supply operation must be tested to determine whether the proper voltages are being supplied to the various vacuum-tube circuits. A careful mechanical examination is also necessary, not only to locate possible faulty construction, but to determine accessibility of parts. This latter is important from the point of view of ease of servicing in the field.

There are numerous detailed points which must be investigated, but since the points needing test differ in each receiving set, it is impossible to set down any general rules. The experience of the test man must be called into use in determining the important special features for test.

Interpretation of Measurements. After measurements have been made and the graphs drawn, the interpretation of the results can be made and conclusions drawn as to the performance of the receiving set. This often requires considerable experience on the part of the engineer. The only way to acquire such experience is to make measurements on a number of receiving sets of various capabilities, and then to compare the data obtained with the degree of practical success which these receiving sets have had under broadcasting conditions in various locations. The relations between electrical measurement data and practical operating results are clearly evident after several comparisons of this sort.

It is important that the tester should realize the necessity of having all the conditions of test such that no anomalous conditions shall be introduced. A forcible example of the effect of such a condition is shown in the fidelity curves of Fig. 12. The solid curve shows the fidelity characteristic of a certain a.f. amplifier under normal conditions. The dashed curve shows the same amplifier with all conditions identical, except that high resistance *B* batteries were used. Space does not permit of further examples of this sort, but frequent cases of this kind are found during tests on receiving sets, and it is important that the tester be sufficiently skilled to recognize and avoid such unreliable test conditions or results. One of the safest methods is to check results by some other method whenever an apparently illogical result is obtained. In this connection, the necessity for standardization in test circuit constants is well illustrated in the graphs of Fig. 13. Here the difference between the two fidelity characteristics is caused merely by a change from a pure resistance output load to a loudspeaker output load.

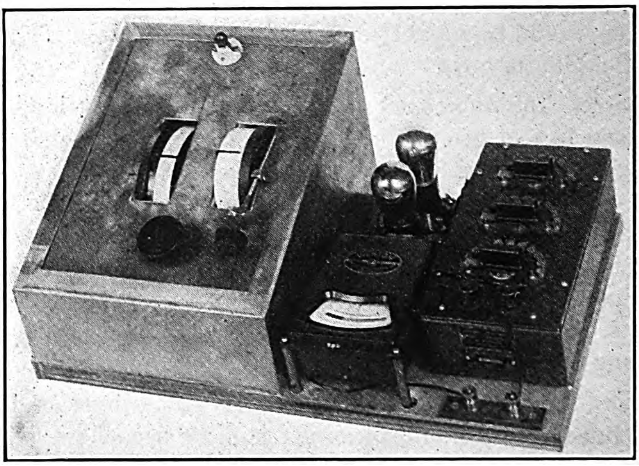

Fig. 18—Rectifier and Meter of R. F. Oscillator.

PRODUCTION TEST EQUIPMENT

As stated previously, the measurement equipment used for the test of receiving sets in large quantities may be simplified in many respects from that used in special engineering tests.

Ruggedness and simplicity of operation are more important, and high accuracy of measurement is less important, than in special engineering tests.

Audio-Frequency Source. For the sensitivity test, a "hummer" signal is used to modulate the r.f. oscillator. For fidelity test, the r.f. oscillator is modulated with phonograph music, and the fidelity is judged by ear, using a standard loudspeaker of a type intended for use with the receiving set under test. The phonograph records for this test are selected with the idea of showing the low- and high-frequency response of the receiving set circuit. An electric pickup is used on the record, and the voltage from it is used to modulate the r.f. oscillator.

Radio-Frequency Oscillator. In the early part of 1926, Dr. Walter Van B. Roberts developed a radio-frequency oscillator circuit which is particularly well adapted to production test work. It has outstanding characteristics of simplicity of construction and operation, minimum necessary shielding, ruggedness of output measuring device, and reliability of operation. The theory on which the functioning of this oscillator is based is briefly described below.

The oscillator is divided into two main portions. One portion consists of oscillator and modulator circuits, and the second portion consists of coupling means for varying the output, a double-wave rectification system, and a d.c. microammeter for measuring the output. The r.f. oscillator operates at a frequency one-half that of the desired broadcast frequency. Fig. 14 shows the basic circuit arrangement. Circuit I is tuned to the fundamental of the r.f. oscillator in order to eliminate the second harmonic of the oscillator from the rectifier circuit. Circuit II consists of two rectifier tubes giving full-wave rectification. Referring to Fig. 15, the action of the rectifiers on the oscillator fundamental frequency will be seen. At *A* is shown the output of the r.f. oscillator, which is assumed to be sinusoidal after it has passed through tuned circuit I. If the double-wave rectifiers worked on a linear rectification principle, the resultant wave form in the rectifier circuit would be as shown at *B* in Fig. 15. However, if the rectifier tubes are worked sufficiently far down on their characteristic curves, they will give practically square law rectification. Therefore the rectified wave will look approximately as shown at *C*. Two interesting facts will be noted with regard to this rectified current. First, it is of double the frequency of the oscillator fundamental; second, it is a sine wave whose axis represents the value of the d.c. component of the rectified current which will be measured by a d.c. meter in series with the rectifiers. This direct current is equal in value to the peak value of the a.c. component of the rectified current. Thus it is possible to measure the a.c. output of this oscillator system by means of a d.c. meter. The advantages of such a system are too obvious to need comment. The only requirements affecting equality of the direct current and a.c. peak values are: (1) That the rectifiers be so adjusted that with no impressed a.c. voltage the current is zero, but that the slightest impressed voltage will cause a current to flow; (2) That the rectified currents be sufficiently small so that the rectifier tubes obey a square law; and (3) That the currents due to the second harmonic of the oscillator are reduced to a negligible value in the rectifier circuit. Those desiring further information regarding the theory of this oscillator system are referred to a paper which has been published describing it.[3]

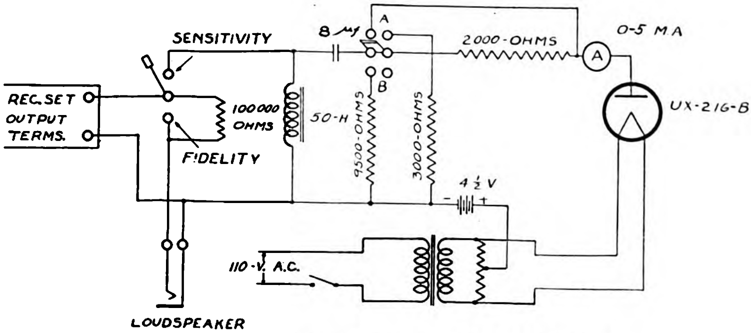

Fig. 19—Production Test Output Meter Circuit.

The complete circuit of this oscillator is shown in Fig. 16. The Heising system of modulation is used in the oscillator. It is necessary to use separate filament supply for the oscillator and rectifier tubes. All r.f. coil systems are astatically wound.

Fig. 17 shows a photograph of the complete oscillator, together with the phonograph system for phonograph modulation.

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[3] Walter Van B. Roberts, "A Method for Generating and Measuring Very Weak Radio-Frequency Currents," *Journal* of the Franklin Institute, March, 1926.

The oscillator and modulator circuit is contained in the sloping panel metal box on the left of the table, and the coupler and rectifier system is at the right. The coupling coil shown at the right of the oscillator table is used in testing loop receiving sets. Fig. 18 shows a closer view of the coupler unit, and of the rectifier and output meter system. The intermediate tuned circuit is controlled by means of a condenser in the coupler shielded compartment. The fact that a very high degree of shielding is not necessary is illustrated by the large openings provided in the sloping panel shield, for the condenser and coupling drums shown in Fig. 18.

Transfer Circuit. For tests on receiving sets using antenna and ground connections, a mutual inductance coupler attenuator is used between the oscillator output and the artificial antenna system. In testing loop receivers, a signal generating coil is connected to the output of the oscillator, as in the case of the engineering test procedure described previously.

Output Circuit. In order to make the Production Test output meter as rugged as possible, it is desirable to have it consist of a rectifier tube and a d.c. milliammeter. Since such a circuit offers a definite load to the circuit across which it is connected, it is made part of the load resistance. By proper selection of series and shunt resistors it is possible to produce a circuit which has the proper load characteristics for either a 2000-ohm or a 5000-ohm output tube. Such a circuit is shown in Fig. 19. By throwing the double-pole switch to position *A*, the circuit simulates a 2000-ohm load, and in position *B* a 5000-ohm load. In either position, normal output (50 milli-watts) is given by a reading of 1 milliampere on the d.c. meter.

As in the case of special engineering tests, a monitoring loudspeaker is provided. By throwing the single-pole switch downward, the loudspeaker is connected directly across the receiving set output for the fidelity listening test.

PRODUCTION TEST CONDITIONS

Test Booth. With open antenna-type receiving sets it is not necessary to use a shielded booth when using the Roberts oscillator, unless the receiving set has very high sensitivity. For loop antenna or highly sensitive open antenna receiving sets, however, a booth is needed. The test booth used for production test work at this laboratory is shown in Fig. 20. The output meter panel can be seen in the right-hand corner of the booth, through the open door. The heavy construction of the door, together with the weather stripping around the door, can also be seen. The upper half of this booth is enclosed with wire mesh. The oscillator is located in the center of the booth. This permits measurement work to be done at one end of the booth while a receiving set is being removed from the other end, and a new set put in and connected up. This arrangement is a considerable aid in speeding up the test work, and makes it possible to use the oscillator more efficiently.

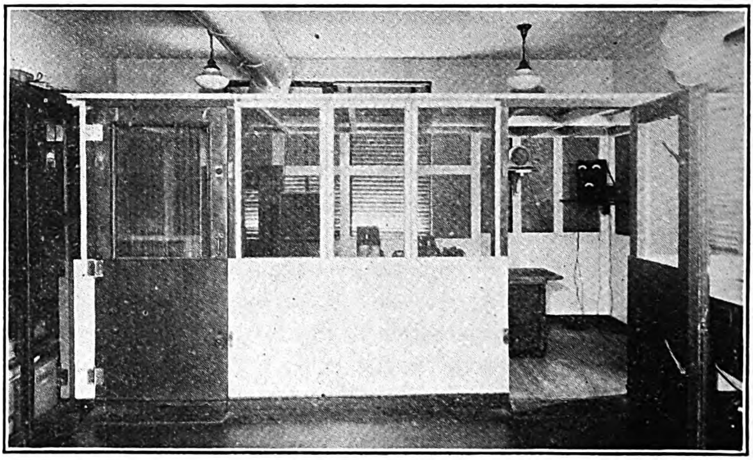

Fig. 20—Production Test Booth.

CONCLUSION

It is the belief of the authors that the day is not far distant when it will be possible for radio engineers to use one universally comprehensible language in speaking of receiving set measurement and performance. It is hoped that this description of test equipment and methods may be of assistance in this direction. In a recent test of the methods, measurements were made on a certain receiving set by four different laboratories, at different times, with results which checked within 25 per cent. Of this error, it has been established that about 10 per cent was due to changes in the characteristics of the receiving set itself, as it was moved from one laboratory to another. This check is the more noteworthy when it is considered that the detail parts of the test equipments in the various laboratories differed in many respects.

The authors wish to acknowledge the assistance which has been rendered by Messrs. Bonanno, Howard, and Whitehead of this laboratory in the design of the test equipment described.

Volume 16, Number 11 *November, 1928*

VACUUM-TUBE PRODUCTION TESTS*

BY

A. F. VAN DYCK AND F. H. ENGEL

(Technical and Test Department, Radio Corporation of America, New York City)

Summary—*General description is given of the methods and apparatus employed by the authors in the testing of vacuum tubes in large quantities.*

EVERY manufacturer is confronted with the problem of determining and controlling the quality of his product. When the number of units produced is small and the product is simple in nature, the problem is relatively simple; when mass production methods are employed on a product of complex nature, and the number of units produced is in the millions, production testing involves difficult problems.

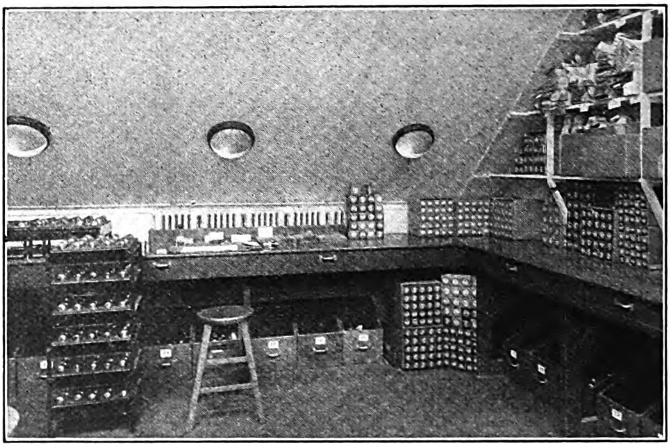

Fig. 1—Receiving Room. Unpacking and Initial Inspection.

Vacuum-tube production has increased rapidly from a total of a few thousands per year, ten years ago, to a yearly production which runs into the millions. The increased demand has made possible the introduction of automatic machinery in many of the manufacturing processes, with the result that vacuum tubes of high quality and uniform characteristics may be had today in any quantity at a fraction of their cost several years ago.

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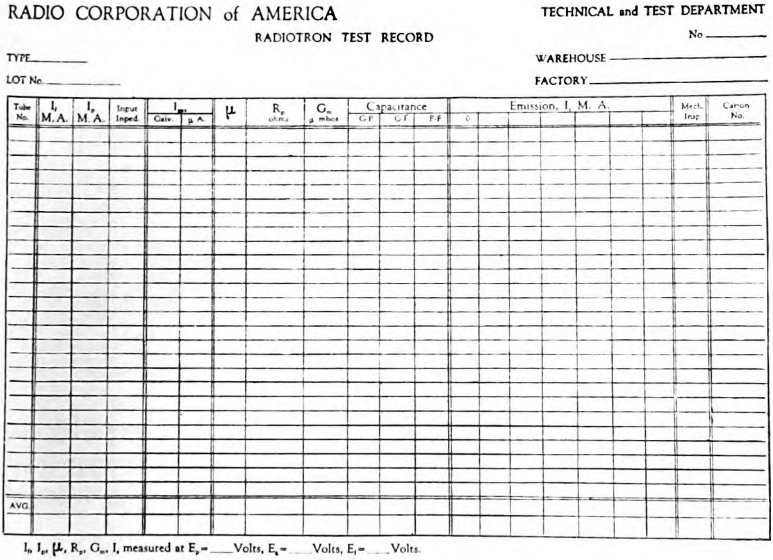
* Original Manuscript Received by the Institute, August 25, 1928. Presented at meetings of the following Institute Sections: Connecticut Valley, September 13, 1928; Rochester, September 14, 1928; Buffalo-Niagara, September 17, 1928; Cleveland, September 18, 1928; Chicago, September 19, 1928; Detroit, September 20, 1928.

In the early days of radio broadcasting the uniformity of the electrical characteristics of vacuum tubes available on the market was relatively unimportant because the circuit requirements were simple and provisions were made in the sets for adjustment of the tube operating conditions. An experienced, skillful operator was able to obtain maximum efficiency in the operation of his set by adjustment of the controls.

Present day multi-tube, single control receivers would be impossible without vacuum tubes having definite, uniform electrical and performance characteristics. Modern circuits with their high degree of selectivity, sensitivity, and fidelity, are critically designed and balanced for use with certain types of tubes of known characteristics. Departure of these characteristics of the tubes from the established values results in inferior performance, and in extreme cases, in complete failure of the set. The satisfaction given by a modern receiving set is definitely dependent upon the degree of uniformity of the characteristics of the tubes employed in the set. This being the case, the importance of maintaining the uniformity and dependability of vacuum tubes is readily appreciated.



RADIO CORPORATION of AMERICA
TECHNICAL and TEST DEPARTMENT
RADIOTRON TEST RECORD
No. ____
TYPE ____
WAREHOUSE ____
LOT No. ____
FACTORY ____

Tube No.	I_f M.A.	I_p M.A.	Input Inped.	I_{gs}		μ	R_p ohms	G_m μ mhos	Capacitance			Emission, I, M. A.							Mech. Insp.	Carton No.
				Galv.	μ A.				G-P	G-F	P-F	0								
AVG.																				

I_f, I_p, μ, R_p, G_m, I_e measured at E_p= ____ Volts, E_c= ____ Volts, E_f= ____ Volts.
I_{gs} measured at I_p= ____ MA, E_p= ____ Volts, E_c= ____ Volts.

Fig. 2—Measurement Record Card

In this paper will be given a general survey of the methods and equipment employed by the Radio Corporation of America in the testing of Radiotrons. The complexity of the equipment and the large number of highly specialized test circuits employed preclude their detailed description at this time. The circuits employed are generally known, and the special forms in which they are used in these tests are determined by practical considerations such as ease of operation, and labor and time saving qualities.

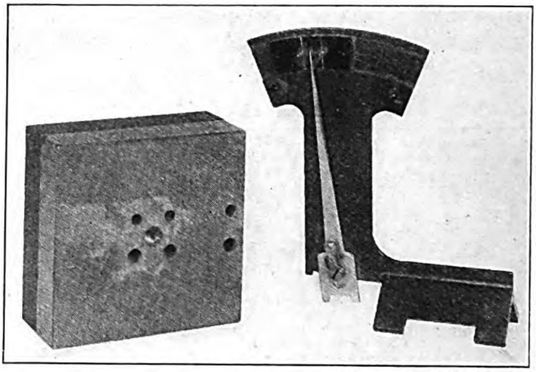


Fig. 3—Typical Gauges Used in Mechanical Inspection.

The vacuum-tube production test methods employed by the Radio Corporation of America may be divided for convenient discussion, into four operations which are listed in the order of their execution.

1. Sampling the product.
2. Initial Inspection.
3. Measurement of initial electrical characteristics.
4. Life Testing.

These operations will be discussed individually.

SAMPLING THE PRODUCT

Great care is necessary in the selection of test samples to insure that the samples selected truly represent the quality of the entire lot of tubes from which they have been selected. The number of tubes selected should not be too small, yet an unnecessarily large number of samples should not be taken for economic reasons. In general, the larger the number of tubes under consideration the smaller the percentage of samples can be. For example, one tube is not safely representative of a group of 100 tubes if 100 is the total number under consideration, whereas one tube might safely be considered representative of 100 tubes where the total being considered runs into the thousands. The percentage of tubes selected for test samples also depends upon the type of tube under consideration and its previous production history.

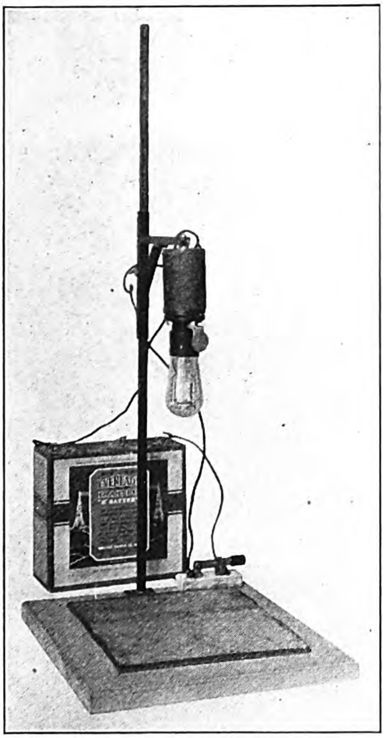


Fig. 4—Drop-Test Apparatus

Sampling of the product is done at the factories, the inspector on duty selecting the finished product at the packing station according to a schedule determined by the rate of production and type of tube. These samples are carefully marked with date of manufacture and other essential information and are forwarded to the testing laboratory.

Samples for test purposes are also taken at intervals from the warehouses to check the results of tests made on samples selected at the factories.

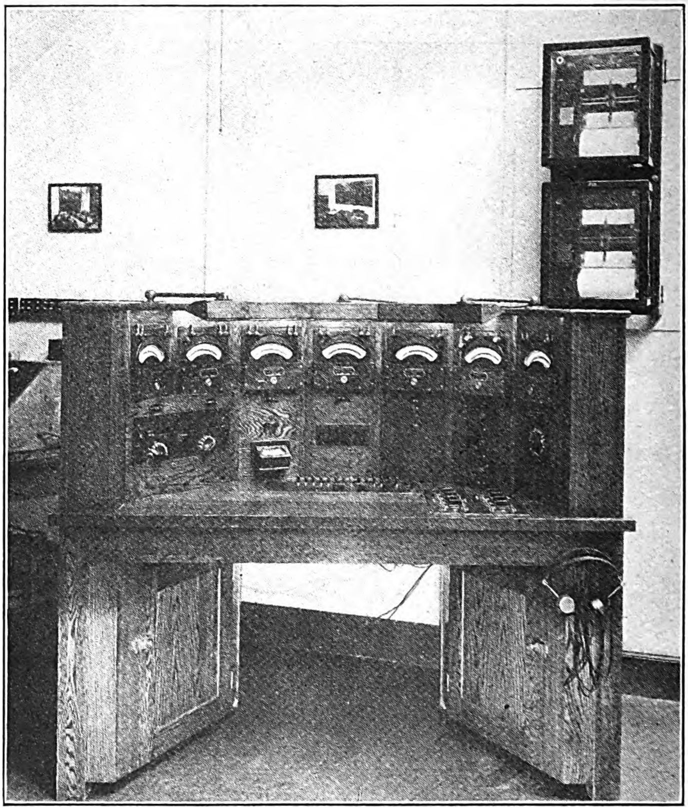


Fig. 5—Universal Test Bridge

INITIAL INSPECTION

Initial inspection of the samples is made at the time the tubes are unpacked at the testing laboratory. Each sample (including carton, wrapper and instruction sheet) is carefully inspected for appearance and mechanical condition. The tubes are labeled with serial number, factory number, and date of manufacture. Fig. 1

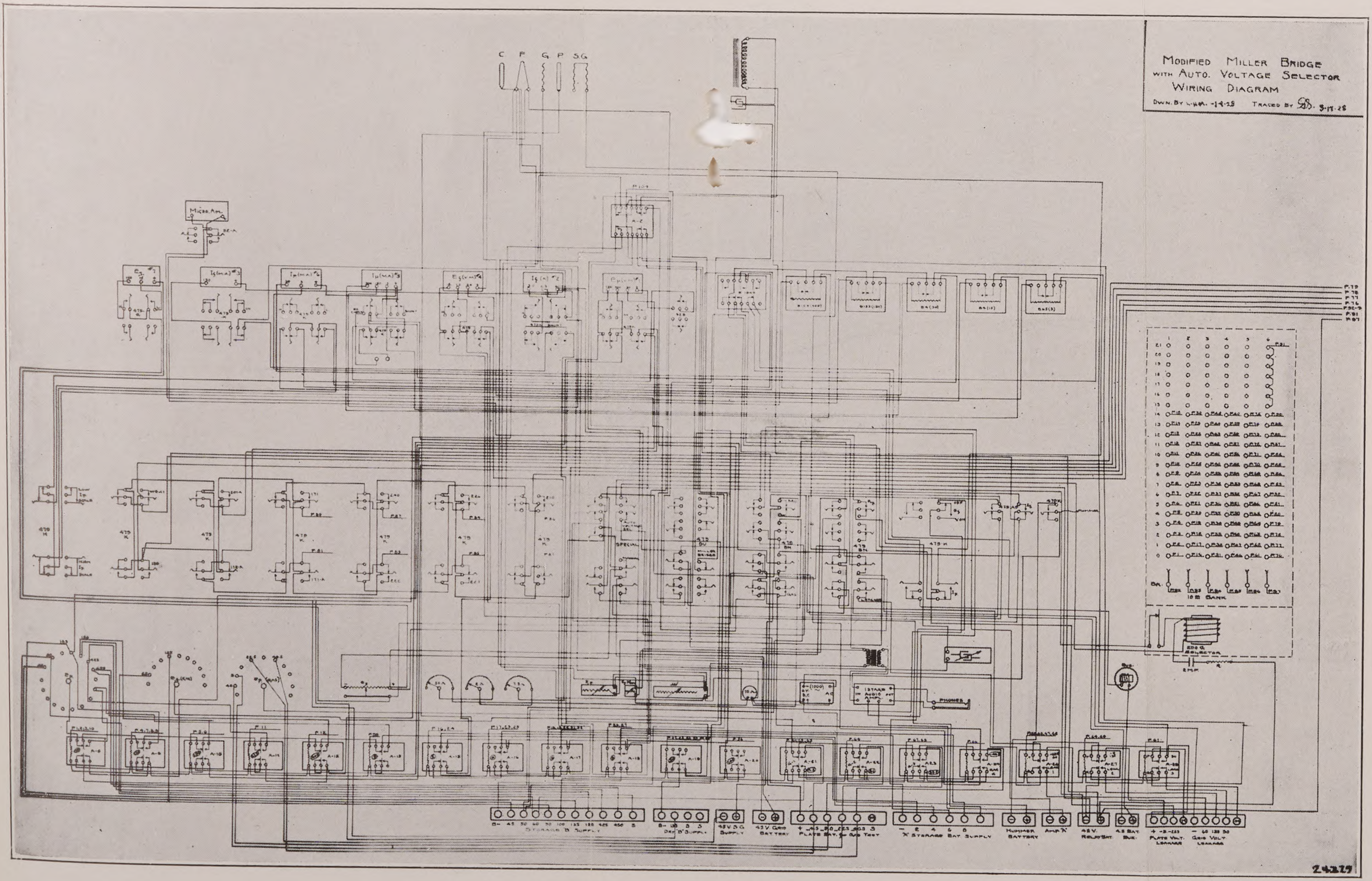


Fig. 6—Wiring Diagram of Universal Test Apparatus.

For certain tests in the mechanical inspection work, special gauges have been provided which make easy the detection of mechanical faults. In Fig. 3 are shown several gauges used in this work. The device on the left is used for determining whether or not excessive solder is on the contact pins; the device on the right is used in measuring the degree of eccentricity of the tube bulb with respect to its base.

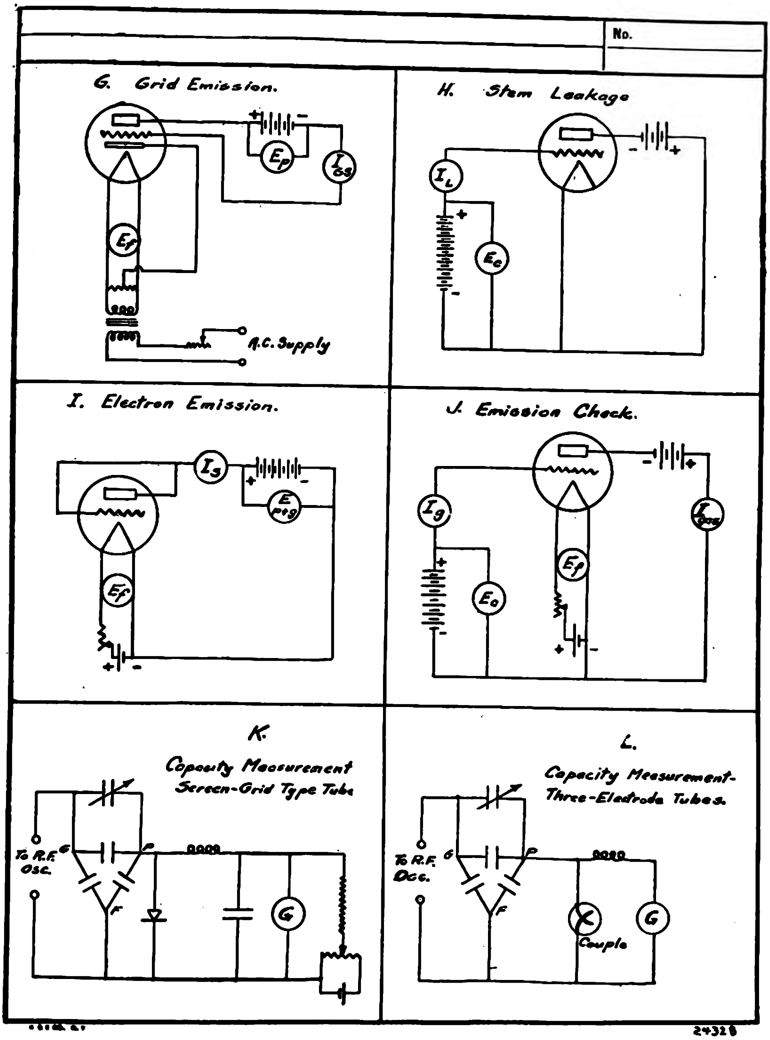


Fig. 8—Typical Test Positions.

Mechanical strength of the product is checked by subjecting the samples to standardized "drop" and "bump" tests. In the "drop" test the tubes are dropped from various heights, and in various positions relative to the base of the test machine. In the "bump" tests the tubes are made the "bob" of a pendulum and allowed to strike a surface after having traveled through a given degree of arc. These tests may be made with or without operating voltages applied to the tubes. Other tests of mechanical strength are made periodically by actual shipment of samples

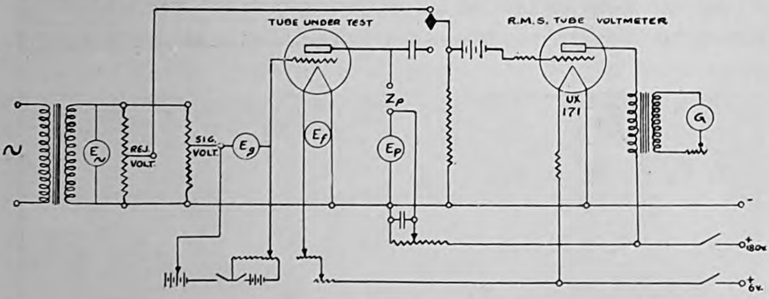


Fig. 9—Universal Test Set. Schematic Diagram.

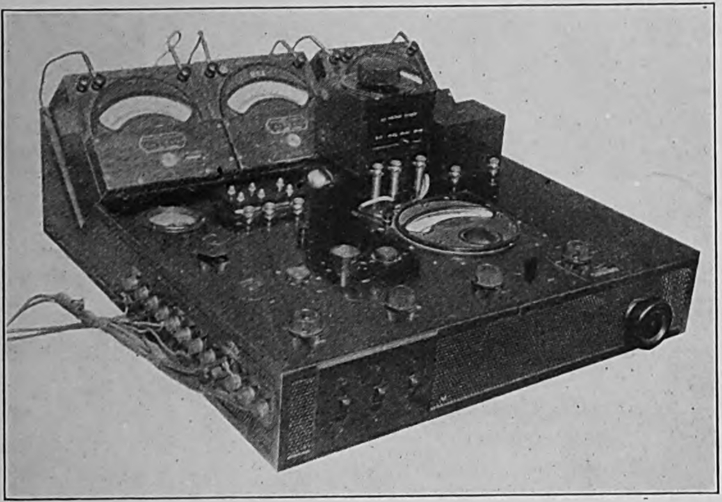


Fig. 10—Universal A.C. Test Set.

about the country under various conditions of packing. One type of drop testing apparatus is shown in Fig. 4.

INITIAL ELECTRICAL CHARACTERISTICS

After being inspected for appearance and mechanical condition, each sample is tested for initial electrical characteristics.

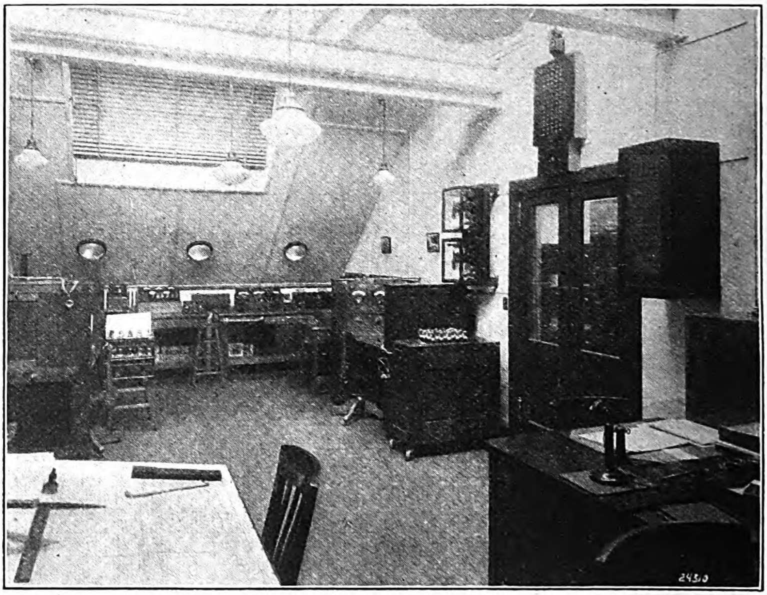


Fig. 11—Characteristic Measurement Laboratory.

The following characteristics are measured, although not every characteristic listed is tested for in every type tube.

1. Filament current at rated voltage.
2. Plate current.
3. Amplification factor.
4. Plate resistance.
5. Mutual conductance.
6. Electron emission.
7. Inter-element (stem) leakage.
8. Gas content (degree of vacuum).
9. Input impedance.
10. Grid emission.
11. Back emission.

12. A.C. test (dynamic).
13. Inter-electrode capacitance.
14. Receiving set performance.

The first ten characteristics listed above are measured with the apparatus shown in Fig. 5. This device is a universal test equipment which, by means of switches, may readily be changed from one circuit condition to another to meet the requirements of the particular test desired. A wiring diagram of this bridge is shown in Fig. 6. One model of this bridge which has been recently developed is automatically adjusted to any desired test position by means of buttons controlling an automatic selector switch. Much time is saved for the operator by the bridge construction, since twenty types of tubes are tested in this bridge and each tube has its own peculiar test circuit requirements.

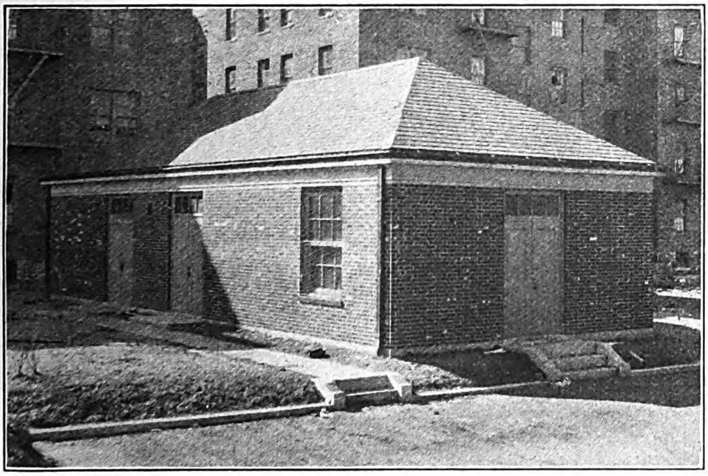


Fig. 12—Power House—Exterior View.

Figs. 7 and 8 give some of the more generally used test positions which it is possible to obtain on this bridge. The circuits shown are the ones generally used, and are conventional in design.

The a.c. test listed above is not generally known and deserves special comment. A circuit diagram of the device is shown in Fig. 9, and an external view is shown in Fig. 10.

The a.c. test is a "dynamic" test, i.e., it measures the overall amplification of a tube under conditions approximating those of service. A definite signal voltage (a.c.) is applied to the input circuit of the tube, and the resulting output signal voltage is measured in the plate or output circuit. The set is flexible and it may be readily changed to test any type of tube by simply inserting the proper potential dividers and load impedances in the jacks provided for that purpose. Much work has been done on the correlation of the indications of this bridge and the actual receiving set performance of tubes, and values have been obtained for the a.c. test for many types of tubes which represent performance limits. In other words, it is possible to predict the behavior

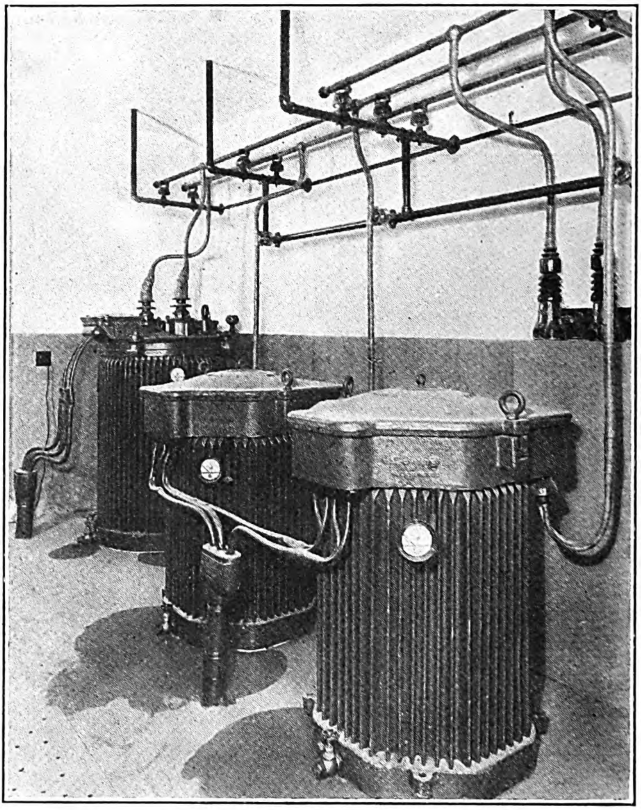


Fig. 13—Power House—High Tension Vault.

of a tube in a receiving set if its a.c. test reading is known. This test is gradually being adopted for all three-electrode tubes as a criterion for the end of useful life during life tests.

In making measurements of the electrical characteristics of tubes, great care in the selection of circuits employed is very important, as are also the construction of the apparatus, calibration of meters, and accuracy with which the readings are taken. Minor details which may seem insignificant, are quite often the cause of serious errors in measurement and inability to check results. A general view of the measurement room is shown in Fig. 11.

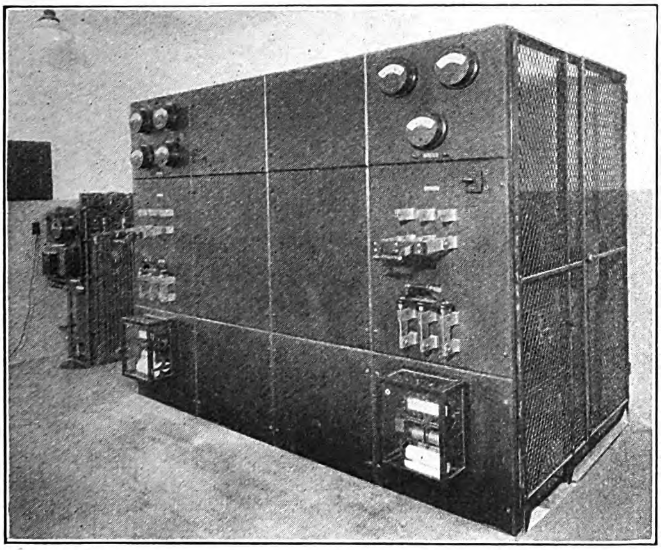


Fig. 14—Power House—Low Tension Vault.

LIFE TESTING

The life quality of vacuum tubes is a very important characteristic and it is also, perhaps, the most difficult one to determine accurately. Absolute control of every factor affecting life performance is essential for accurate results, because apparently insignificant details of the test apparatus or method may have enormous influence on the test results.[1]

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[1] W. C. White, "Life Testing of Tungsten Filament Triodes" PROC. I.R.E., 13, 625; October, 1925.

The test equipment described represents the accumulated knowledge gained from testing thousands of tubes during the past few years. Every precaution is taken to insure the reliability of the results obtained. The tests are operated continuously throughout the year so that judgments of life quality of any type of tube are not restricted to the results of a few tests but may be based on a large number of tests made on samples representative of the entire production.

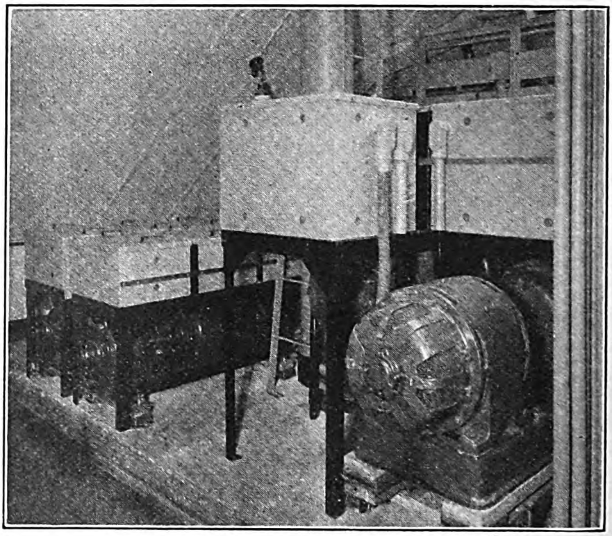

Fig. 15—Motor-Generators for D.C. Power Supply.

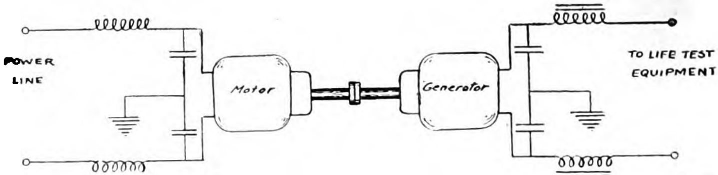

Fig. 16—Motor-Generator Filter Circuits.

POWER SUPPLY

The power used in life testing Radiotrons (New York City Laboratory) is obtained from the local power company. This power is transformed and regulated (automatic voltage regulators) in a separate substation building. Fig. 12 is an exterior

Grid voltage (d.c.) is supplied by batteries of dry cells, the load on this circuit being quite small and economically handled by the No. 6 type dry cell.

Direct-current power for use in making initial bridge tests and all intermediate measurements (during life) is obtained from storage batteries. The high-voltage battery equipment and charging panel are shown in Fig. 18.

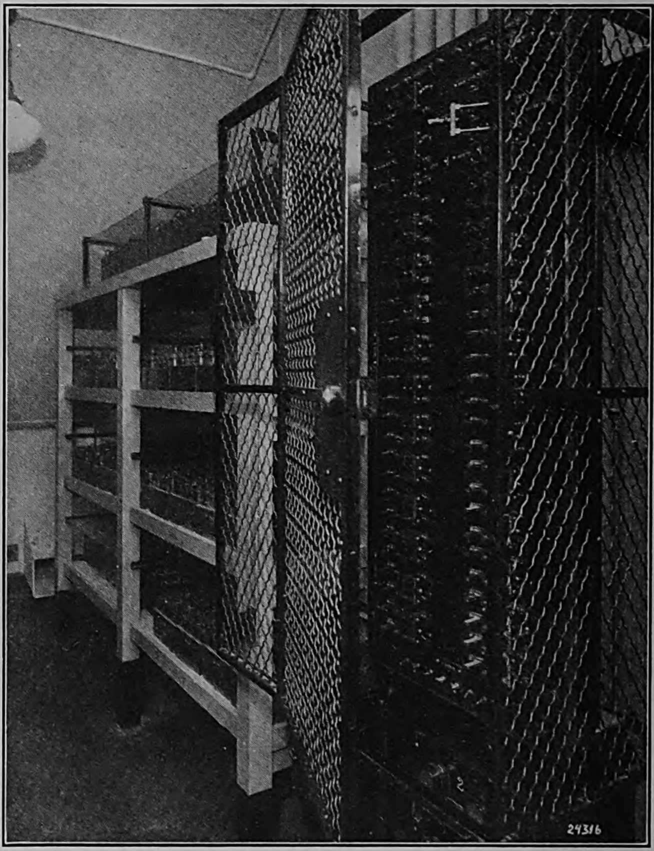

Fig. 18—High-Voltage Battery and Charging Equipment.

A large portion of the power used in life testing vacuum tubes is dissipated in the form of heat. Because of the large number of tubes tested in this laboratory, special precautions are necessary to control the temperature of the life test room. A cooling system

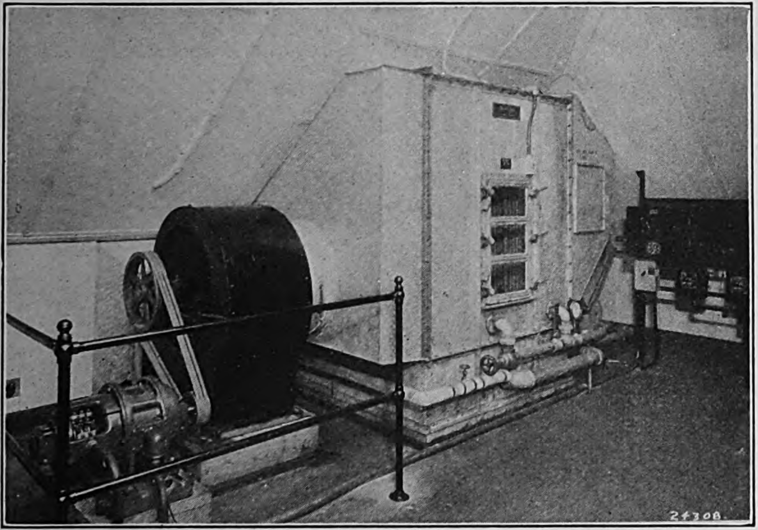

Fig. 19—One Unit of Cooling System.

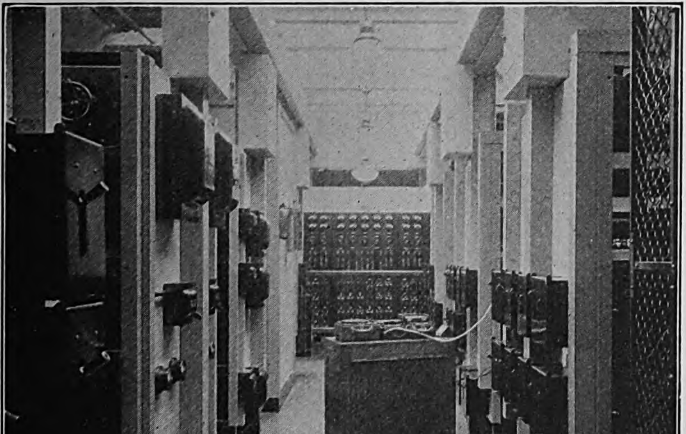

Fig. 20—Life Test Rack Room.

Series non-inductive resistors are placed in the grid circuits of all tubes to prevent oscillation of the tubes, and are of such value that they serve also to prevent removal of the grid bias potential from the rest of the tubes on the rack when one tube on test fails because of grid to filament short circuit.

Plate power is supplied to the tubes in banks of seven tubes, each bank having a "grasshopper" fuse in circuit which opens that particular circuit if abnormal conditions occur. A balanced relay arrangement is also included in the main plate power supply line of each shelf of tubes which automatically shuts off the filament and plate power to that shelf if the grid, filament or plate power supply should fail, or vary more than a specified amount from normal. An alarm is also provided and the location of trouble is indicated on a signal board located in a prominent position in the test room. The relay panels may be seen in Fig. 21. The indicator panel is shown in Fig. 22.

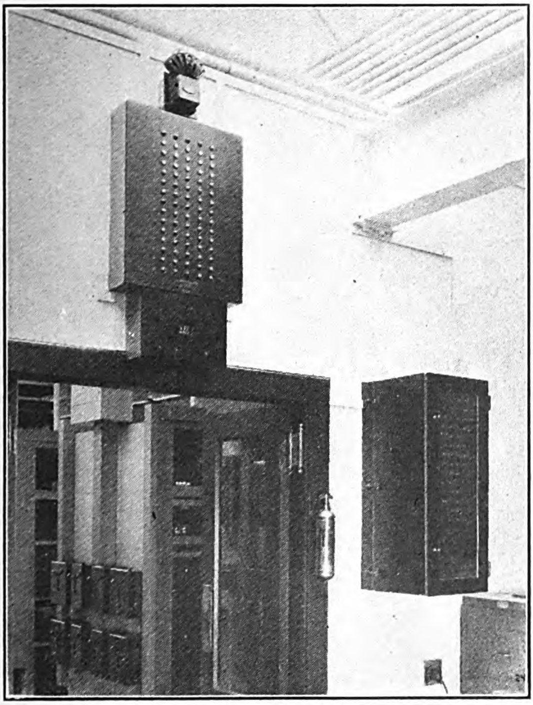

Fig. 22—Alarm Indicator and Timing Devices.

the racks are in normal operating condition the impulse is passed on to the relay counters (telephone message counters) shown in Fig. 22. A counter-relay is provided for each shelf of each rack. If a particular shelf is shut down, the timing impulse does not operate the counter-relay connected to that shelf, so that by observation of the totals shown on the relay dials the amount of time a particular shelf has been shut down is readily ascertained. Since all racks are automatically controlled and may shut down when an operator is not present (during the night) this timing system is necessary in order to obtain accurate timing of the tests.

Both intermittent and continuous burning life tests are conducted, the kind of test used depending upon the type of tube being tested. The continuous test is preferred because no time is lost as contrasted with intermittent testing which may utilize as little as half of the test time for actual operation of the tubes. Most types of receiving tubes give equally good results on either type test, so that most of the tests are of the continuous type.

During life test, the condition of the tubes is determined by measurement at regular intervals. For example a typical test schedule is 0–10 per cent–20 per cent–40 per cent–75 per cent–100 per cent of the total run of hours. The tubes may be tested at any intermediate time at the discretion of the operator, but the regular intervals are preferred for the sake of uniformity in the test reports.

Volume 16, Number 11 *November, 1928*

Supplementary Note† to ABBREVIATED METHOD FOR CALCULATING THE INDUCTANCE OF IRREGULAR PLANE POLYGONS OF ROUND WIRE*

By

V. I. Bashenoff

(Chief of the Wireless Section of the State Research Institute, Moscow, Union of Socialist Soviet Republics)

The author has made an analysis of the nature of the quantity a_k which occurs in the principal formulas of the foregoing paper.*

The formula (11) for the inductance of a rectangle of sides a, b, diagonal d, perimeter $l=2(a+b)$ may be transformed by means of the relation

$$L = 2l\left\{\frac{2(a+b)}{l}\log\frac{ab}{l^2}\cdot\frac{2l}{r}\cdot l - \frac{2a}{l}\log(a+d) - \frac{2b}{l}\log(b+d) + \frac{2(a+b)}{l}\mu\delta + \frac{4d}{l} + \frac{4r}{l} - \frac{4a}{l} - \frac{4b}{l}\right\} \tag{44}$$

$$= 2l\left\{\log\frac{2l}{r} + \log\frac{ab}{l^2} - \frac{2a}{l}\log\frac{(a+d)}{l} - \frac{2b}{l}\log\frac{(b+d)}{l} + \frac{4d}{l} - 2 + \mu\delta + \frac{4r}{l}\right\}$$

into the form

$$L = 2l\left\{\log\frac{2l}{r} - a_k + \mu\delta + \frac{4r}{l}\right\} \tag{45}$$

where

$$a_k = -\log\frac{ab}{l^2} + \frac{2a}{l}\log\frac{a+d}{l} + \frac{2b}{l}\log\frac{b+d}{l} - \frac{4d}{l} + 2 \tag{46}$$

Calculating a_k for different values of a/b, and therefore for different values of $l/\sqrt{s}$, the following table is obtained.

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* Proc. I.R.E. 15, 1013; December, 1927. The numbers of equations, tables, and figures here follow those of the previous paper.

† Original Manuscript Received by the Institute, July 4 1928. Translated from the German by F. W. Grover, Union College, Schenectady, N. Y.

TABLE V

a/b	1	2/3	0.427	1/4	1/9	1/15	1/20	1/30	1/50
$l/\sqrt{s}$	4	4.08	4.37	5	6.67	8.24	9.35	11.3	14.37
a_k	2.853	2.866	3.006	3.269	3.826	4.226	4.471	4.849	5.330
$\log l^2/s$	2.773	2.79	2.94	3.22	3.80	4.218	4.471	4.849	5.330
ϕ	0.080	0.076	0.066	0.049	0.026	0.008	0	0	0

The quantity ϕ is put for $a_k - \log l^2/s$.

It is clear that in (46) the first term $-\log ab/l^2 = \log s/l^2$ is the most important, and is increasingly so, the greater the value of $l/\sqrt{s}$ (see last row of Table V).

Replacing a_k by $\log(l^2/s)+\phi$ in (45) we obtain

$$L=2l\left[\log\frac{2s}{rl}+\mu\delta+4\frac{r}{l}-\phi\right] \tag{47}$$

and if we neglect here, as has been done earlier, terms of the order of magnitude of r/l and ϕ,

$$L \leqq 2l\left[\log\frac{2s}{rl}+\mu\delta\right] \tag{48}$$

From this may be obtained, as was first suggested to me by Ing. Klatzkin, the known formula for a long loop[1]

$$L \leqq 2l\left[\log\frac{d}{r}+\mu\delta\right] \tag{49}$$

The assumption will now be made that for other closed curves without reëntrant angles the quantity a_k is likewise nearly equal to $\log(l^2/s)$.

To prove this assumption analytically is, unfortunately, possible only in the two cases of a rectangle and a triangle, using the formulas derived by the author and Dr. Grover. Besides these, no other cases exist where, for a closed curve, formulas with two variable parameters can be given in analytical form.

By a process similar to that already described for a rectangle, formula (42) for a right triangle may be brought into the form

$$L= 2l\left[\log\frac{2s}{rl}+\mu\delta-\phi\right] \tag{51}$$

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[1] More exactly, if l_1 is the length of each wire, and d their distance apart,

$$L=(4l_1+4d_1)\left\{\log\frac{dl_1}{r(l_1+d)}+\mu\delta\right\} \tag{50}$$

where

$$\phi = -\frac{a^2}{2lc}\log\frac{a(c-a)}{b(c+b)} - \frac{b^2}{2lc}\log\frac{b(c-b)}{a(c+a)} - \frac{c}{2l}\log\frac{c^2}{ab}$$

$$+\frac{a+c/2}{l}\log(a+c) + \frac{b^2+c/2}{l}\log(b+c) - \log l + 1 - \log 2 \qquad (52)$$

From this follow the results of Table VI.

TABLE VI

$l/\sqrt{s}$	4.828	5.84	7.01	8.99	10.2	12
ϕ	0.182	0.223	0.252	0.271	0.276	0.282

The formula (41) for an isosceles triangle may also be written in the form (51) with

$$\phi = 1 - \frac{2a}{l}\log\frac{2a}{h} - \frac{c}{l}\log\frac{8h}{l} \qquad (53)$$

Here h is the altitude of the triangle.

This last formula gives the results shown in Table VII.

TABLE VII

$l/\sqrt{s}$	4.559	5.67	6.94	8.01	8.95	9.8	10.59	11.3	12	12.65
ϕ	0.164	0.225	0.260	0.276	0.285	0.290	0.293	0.296	0.298	0.299

Furthermore, all the previously known formulas for polygons (32) to (37), (14) and (37a) can also be put into the form of (51). Table VIII gives the data for each.

TABLE VIII

Figure	$l/\sqrt{s}$	a_k	New Formula	$\log l^2/s$	ϕ
Circle	3.541	2.451	$L=4\pi a\left[\log\frac{a}{r}-\phi\right]$	2.5311	−0.080
Octagon	3.641	2.561	$L=2na\left[\log\frac{K_n}{r}-\phi\right]$	2.5845	−0.0235
Hexagon	3.722	2.636		2.6285	+0.0065
Pentagon	3.812	2.712		2.6763	+0.0355
Square	4.000	2.852	$L=8a\left[\log\frac{a}{2r}-\phi\right]$	2.7730	+0.0800
Figure 3a	4.395	3.091	$L=2l\left[\log\frac{2s}{rl}-\phi\right]$	2.96095	+0.130
Equilat. Triangle	4.559	3.198	$L=6a\left[\log\frac{a}{2\sqrt{3}r}-\phi\right]$	3.0340	+0.164
Right Triangle Equal legs	4.828	3.331	$L=(2+\sqrt{2})a\left[\log\frac{a}{(2+\sqrt{2})r}-\phi\right]$	3.1489	+0.182

From this it is clear that for all the regular polygons we can write a new simpler formula

$$L=2na\left[\log\frac{R_u\cos\dfrac{\pi}{n}}{r}-\phi\right]=2na\left[\log\frac{K_n}{r}-\phi\right]$$

in which n = the number of sides.

R_u = the radius of the circumscribed circle.

K_n = the apothegm of the regular polygon.

In the various cases ϕ varies between the limits of -0.08 for the circle and $+0.164$ for the equilateral triangle.

The similar relation is not difficult to derive for other figures with a known ratio of l and s. For example, for any triangle

$$L=2l\left[\log\frac{R_e}{r}-\phi\right]$$

where R_e is the radius of the inscribed circle, and ϕ is taken from the curve, Fig. 1 (see below).

For a rhombus whose shorter diagonal is h and whose acute angle is 2α, the corresponding formula is

$$L=2l\left[\log\frac{l\sin 2\alpha}{8r}-\phi\right]=\frac{4h}{\sin\alpha}\left[\log\frac{h\cos\alpha}{2r}-\phi\right]$$

The curves $\phi=f(l/\sqrt{s})$ for Tables V, VI, VII, and VIII are shown in Fig. 1. For figures of the form of a rectangle ϕ approaches zero with increasing $l/\sqrt{s}$: for polygons more nearly of the form of a right triangle or isosceles triangle ϕ increases, but relatively quickly attains, with $l/\sqrt{s}=10$, a nearly constant

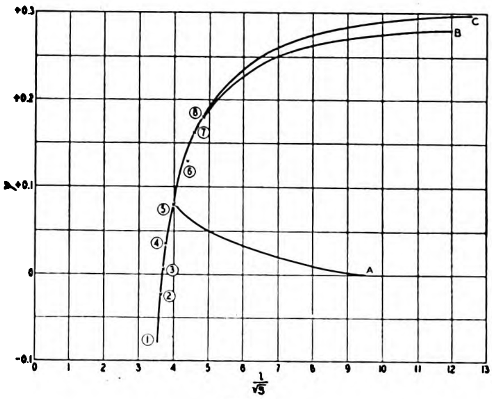

Fig. 1

value of about 0.29. For all practical cases ϕ can be determined from the curves. If, however, the figure is strongly unlike that of a triangle (Curves B and C) or that of a rectangle (Curve A) the author proposes for practical purposes to use the formula

$$L \leqq 2l\left[\log\frac{2s}{rl}-0.15\right]$$

The maximum error for the most unfavorable cases amounts with this formula to only 1.3 per cent.

Thus the new form of our formula, in which the "noninductive effect" of the factor $l/\sqrt{s}$ is more clearly and distinctly expressed, can be used in practice for any value of $l/\sqrt{s}$ without the necessity of extrapolation as was the case previously.

In a recent paper R. G. Allen[2] has obtained the inductance of an equilateral triangle, a rectangle, and a regular hexagon by direct integration of the flux linkages in each of these cases. The resulting formulas agree with those obtained from the general abbreviated formula of the present paper.

At the beginning of March, the author made a measurement of the inductance of an ellipse in order to prove the new formula. The ellipse had semi-axes of 1.5 m and 3.47 m and consisted of wire 0.08 cm in diameter. The lead wires to the bridge were 0.5 m long and were placed 0.2 m apart. The measured value, mean of the results of three observers, was 30800, with an uncertainty of one per cent. The calculated value, assuming a_k to have the value given by Fig. 6 for $l/\sqrt{s}=4.02$, and including the calculated value of the inductance of the lead wires, was 30775. Thus $\Delta L=(Lm-Lc)/Lc=0.08$ per cent.

In view of the difficulty of accurately calculating the capacity of antennas,[3] the author suggests that it is better to obtain the calculated capacity of a closed aerial from its calculated inductance L_0. Supposing, for example, an aerial of a single turn, closed except at a single point, the natural wavelength $\lambda_0=2l$ and the electrostatic capacity C_0 is given by the relation $\lambda_0=4\sqrt{L_0C_0}$, which using the author's inductance formula gives

$$C_0=\frac{l}{8\left[\log\frac{2s}{rl}-\phi+\mu d\right]\left(1+\frac{\Delta}{100}\right)}$$

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[2] *Exper. Wireless and Wireless Eng.*, 5, No. 56, p. 259, Editorial note by G. W. O. Howe, l.c. p. 238.

[3] Proc. I.R.E., 15, 733; August, 1927.

In conclusion the author expresses his sincerest thanks to Prof. Schuleikin and to Electrical Engineers Starik and Svistoff for their interest and aid in the work, and to Prof. F. W. Grover for making the English translation of the paper.

Note added by the translator.

At the request of the author I am adding the general expressions for the quantity ϕ for two further cases.

For a triangle of sides a, b, and c, it may be shown that

$$\phi = 1 + \frac{b+c}{l}\log\frac{(c+b)^2-a^2}{v} + \frac{a+b}{l}\log\frac{(a+b)^2-c^2}{v}$$

$$+\frac{a+c}{l}\log\frac{(a+c)^2-b^2}{v} - \frac{a}{l}\log\frac{2a}{R_e} - \frac{b}{l}\log\frac{2b}{R_e} - \frac{c}{l}\log\frac{2c}{R_e}$$

in which

$$R_e = \sqrt{\frac{(P-a)(P-b)(P-c)}{P}} \quad \text{and} \quad \frac{l^2}{2s} = \frac{a+b+c}{R_e}$$

$$v = 2l\,R_e \qquad P = \frac{l}{2} = \frac{1}{2}(a+b+c).$$

For a rhombus whose acute angle is 2α,

$$\phi = 2 - \log 4 - \cos\alpha - \sin\alpha + \log\sin 2\alpha$$
$$+\cos^2\alpha(\sinh^{-1}\cot 2\alpha + \sinh^{-1}\tan\alpha)$$
$$+\sin^2\alpha(\sinh^{-1}\cot\alpha - \sinh^{-1}\cot 2\alpha)$$

F.W.G.

Volume 16, Number 11 *November, 1928*

THE CONSTANT IMPEDANCE METHOD FOR MEASURING INDUCTANCE OF CHOKE COILS[*]

By

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Summary—*This paper describes a method for measuring at power frequencies the inductance of iron-cored choke coils of the type commonly used in radio as filters. It has been used to measure values of from one to more than 2000 henries. Two circuit arrangements are illustrated and several families of curves show how the inductance depends upon the magnitude of the superposed alternating and direct-current magnetomotive forces. The method may also be used at higher frequencies. It is simple, convenient to use, and requires the minimum of equipment.*

In 1918 the writer discovered a unique property of parallel circuits which was found useful in measuring circuit constants and has been used quite successfully in our laboratory for this purpose for several years.

With an alternating emf impressed upon a parallel circuit as shown in Fig. 1 the equation for the line current is

$$I=EY=E(G+jB)=E[g_2-j(b_2-b_1)] \tag{1}$$

$$=E\sqrt{\left(\frac{R_2}{Z_2}\right)^2+\left(\frac{X_2}{Z_2}-\frac{1}{X_1}\right)^2} \tag{2}$$

$$=E\sqrt{\frac{1-2w^2LC}{R_2{}^2+X_2{}^2}+w^2c^2} \tag{3}$$

It will be observed that if the values of w, L and C are such that

$$2w^2LC=1 \tag{4}$$

the first term under the radical reduces to zero and the absolute value of the line current is equal to wCE, the current through the capacity branch; under these conditions the line current is absolutely independent of the resistance of the inductive branch. This will be referred to as the *critical condition.* Opening the inductive circuit is equivalent to introducing into this branch an infinite resistance for it reduces the inductive current to zero.

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[*] **Original Manuscript Received by the Institute, July 14, 1928.**

It is this feature that is the basis of the method being described. It is unnecessary to obtain a resistance balance as in Wheatstone bridge methods of measuring inductance, nor is it necessary to determine the resistance as in other methods.

Solving (4) gives for the critical frequency

$$f=\frac{1}{\sqrt{2}\,2\pi\sqrt{LC}}=0.707\,f_r \tag{5}$$

when f_r is the resonant frequency for L and C connected in series and

$$L=1/2w^2C \tag{6}$$

When (4) is satisfied and any two of the factors are known, the third may be calculated, but it is usually the inductance that is desired which is expressed in terms of frequency and a standard capacity.

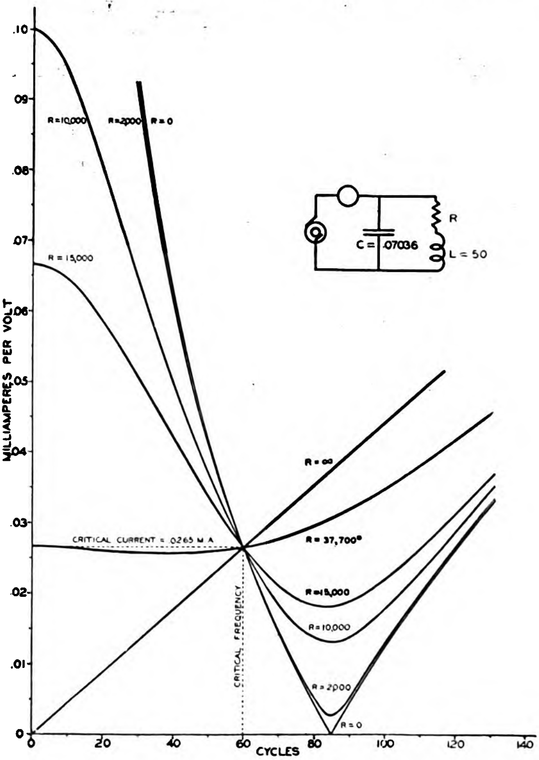

Fig. 1—Line Current Variation in Parallel Circuit Due to Frequency Variation.

For constant impressed emf and critical frequency the resistance of the inductive branch may be varied from zero to infinity without affecting the magnitude of the line current as shown by the common point of intersection of the curves in Fig. 1, while the corresponding change in phase of line current is indicated in Fig. 2. The current in the inductive branch is given by

$$I=\frac{E}{Z}=\frac{E}{X}\sin\theta \tag{7}$$

the equation of a circle which is the locus of the current as R varies. The current for $R=0$, $\theta=90$ deg., is E/X the diameter of the semicircle as given by OA lagging ninety degrees behind the emf. The condenser current is wCE leading the emf by ninety degrees. Adding these two components vectorially gives the line current, the locus of which is a semicircle of radius wCE with its center at the intersection of the coordinate axes. The critical value of capacity current is $E/2X$, that is, the capacity reactance is twice that of the inductive reactance.

PROCEDURE

In general a source of constant emf of known frequency (usually 60 cycles when measuring large choke coils where this method is particularly good), some means of measuring the current, and a standard capacity are required. It should be pointed out that while it is highly desirable to have the voltage remain constant during the entire time measurements are being made, it is only necessary that it be the same immediately after the switch is opened as it was immediately before.

It has been previously stated that when the values of w, L, and C are such as to satisfy (4) the line current has the same value regardless of whether the inductive circuit is connected or not; that is, the total impedance of the circuit is the same, and for this reason has been called the "Constant Impedance Method."

To determine whether the necessary conditions are fulfilled is extremely simple. A switch placed in the inductive branch is opened and closed while the standard capacity is varied until there is no permanent change in the reading of the line ammeter indicating that (4) is satisfied. The inductance may then be calculated by (6). With the circuit connected as shown in Fig. 2, first close the switch and observe the steady ammeter reading;

then open the switch, and if the current decreases the capacity is too small. The capacity is changed in accordance with these observations and the operation repeated until there is no permanent change in the deflection of the instrument when the switch is opened. The transient changes are ignored except insofar as they are useful in assuring the operator that the pointer moves freely as the balance is closely approached. The switching operation may be performed in the reverse order if desired; that is, noting the ammeter reading with the switch open and after closing, but the method outlined is preferred.

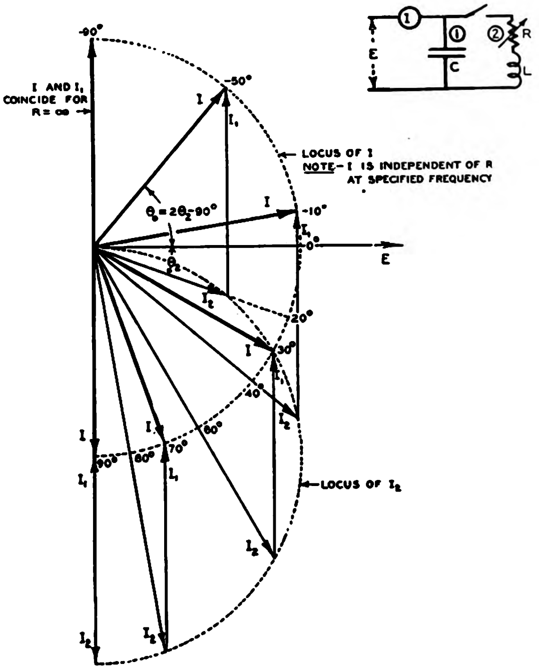

Fig. 2—Change of Phase of Line Current Due to Change in Resistance, with Critical Impressed Frequency.

Under certain conditions the voltage of the source may be disturbed by the change in phase of the line current when the inductive circuit is opened. For example, in Fig. 2, if $R = 0$ the inductive current lags 90 deg. behind the impressed emf. For a condition of balance the line current also lags 90 deg. when the switch is closed and leads by 90 deg. when the switch is open. If R is such that the inductive current lags by 80 deg., the line current lags by approximately 70 deg. when the switch is closed and leads by 90 deg. when the switch is open. This may be avoided by connecting the inductive circuit first on the right of the ammeter and then on the left, thus keeping the magnitude and phase of the line current the same. Also, if the impedance of the line instrument is such as to cause an appreciable voltage drop, the fact that the current is displaced 70 deg. from the voltage in one case and 90 deg. in the other may cause a change in the voltage across the circuit being measured for the two conditions. This increases the voltage across the condenser when the inductive circuit is open; therefore, the capacity must be reduced to give the same line current and the calculated value of inductance is high. This error may be eliminated by increasing the frequency until the inductive components lag by 85 deg. or more. The error introduced by the ammeter impedance is usually not large. The inductance of an air-cored coil as measured with a five-ohm instrument was 4.32 henries and with a one-hundred-and-five-ohm instrument was 4.38.

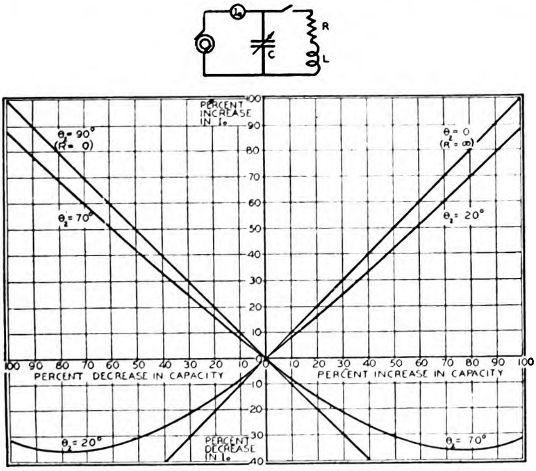

Fig. 3—Measurement of Inductance. Change in Sharpness of Tuning of Parallel Circuit Due to Change in Capacity. Impressed Frequency Critical for $C = 100$ per cent.

Fig. 3 shows the per cent change of line current for a given per cent change in capacity in terms of the departure from the critical value of capacity. The point of intersection of the various curves (zero change) is the critical adjustment where the inductive circuit may be opened and closed without producing any permanent change in the line current. As a balance is approached quite closely, which is always the case in making measurements, the curves are symmetrical with respect to the zero point. It is to be observed that when the angle of lag of the inductive circuit approaches ninety degrees, which is usually the case, the line current changes at the same rate as the capacity and for the case where the angle is small the current changes somewhat less rapidly than the capacity.

It should be observed that with this method, where the apparent inductance of the coil is not affected by the distributed capacity of its winding at the frequency used, higher harmonics of considerable magnitude may be present in the electromotive-force wave without appreciably influencing the calculated inductance. For example, if there is a pronounced 29th harmonic, due to armature slots, the harmonic component of current in the inductive branch is absolutely negligible, while that through the capacity is large. With the switch closed the line current $=\sqrt{I^2+I^2_{c(29)}}$, I being the vector sum of the fundamental components in the two branches, and with the switch open the line current $=\sqrt{I_c^2+I_c^2{}_{(29)}}$ but for balance the fundamental component of the line is equal in magnitude to the fundamental component through the condenser. In other words the fundamental components combine in exactly the same manner as if the harmonic were not present, and since the harmonic component of condenser current is present in the line both with the switch open and closed there is no error introduced. If harmonics of low order but large magnitude are present, error would result because of the relatively large harmonic component through the inductance which would pass through the ammeter with the switch closed but not when open. Either use a wave of approximately sine form or use a filter to eliminate the troublesome harmonic, but usually this will not be necessary.

Iron-Cored Choke Coils

For measuring the inductance of iron-cored coils the procedure is exactly the same as already outlined. However, the inductance will vary with the impressed emf, and this method affords a simple and convenient means of determining the manner in which it depends upon the impressed voltage or current through the coil. Fig. 4 shows a curve of the inductance of a small filter choke plotted against current. Of course, under operating conditions the alternating current would not reach such large values. It was designed to carry fifteen milliamperes of direct current in normal use; however, there was no direct current flowing when these measurements were made.

When operating iron-cored coils at high values of saturation it is extremely important to have the line ammeter short-circuited at the moment the inductive circuit is connected, otherwise the instrument may be damaged by the large transient current. If an extra contact is added to a single-pole switch as shown in Fig. 4, the required protection is provided and the switch may be closed as rapidly as desired without damage to the instrument.

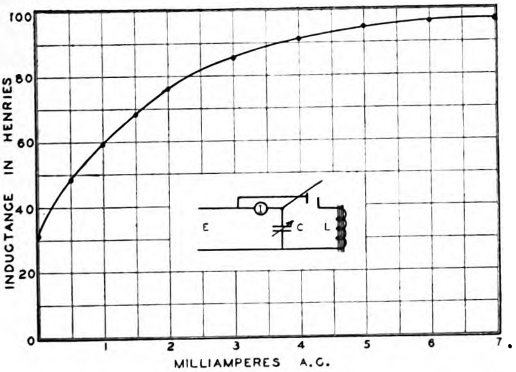

Fig. 4

IRON-CORED CHOKE COILS WITH SUPERPOSED DIRECT CURRENT

Since the inductance of an iron-cored coil varies with the superposed direct current flowing through its winding, provision

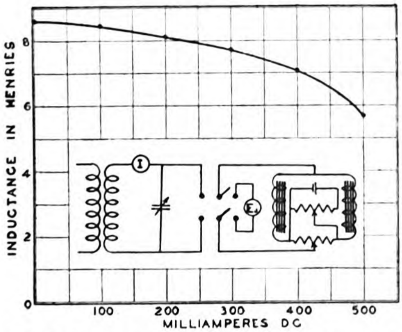

Fig. 5—Variation of Inductance with D.C. Magnetization. Filter Reactor Model UP–1627. 60 Cycles $E_{A.C.} = 110$.

should be made for supplying the desired value of direct current when the inductance is being measured. This necessitates a slight modification in the circuit arrangement. Where two similar coils are available they may be connected as in Fig. 5, providing a convenient method of introducing and controlling the direct current through the coils and preventing battery current from flowing through the thermo-couple instrument in the line. This is accomplished by varying the position of the contact B of the potential divider until there is no battery potential between A and B as indicated by a sensitive voltmeter E_{dc}. This is important for the protection of the line instrument, for a relatively small unbalanced battery voltage might burn it out. When the adjustment is once made it seldom requires changing, but the balance should be checked occasionally by throwing the switch to the right.

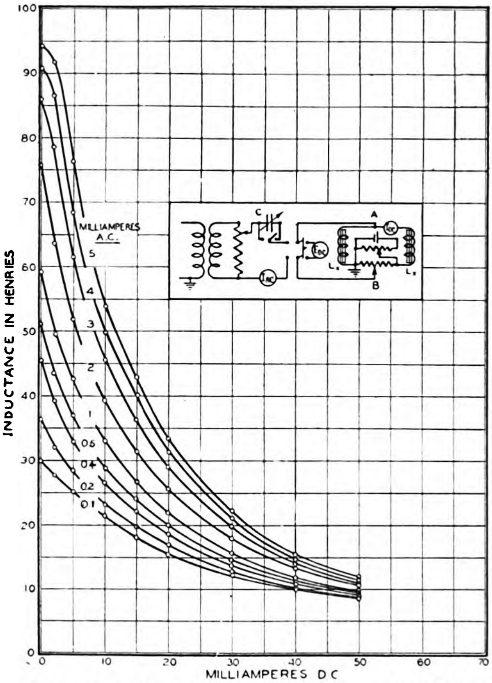

Fig. 6—Variation of Inductance with A.C. and D.C. Magnetization. 60 Cycles.

It is evident that the inductance of the two coils in parallel is measured, and to obtain that of a single coil this value should be doubled.

Fig. 5 shows the inductance of a filter reactor with direct current. The impressed emf was constant and of quite large value, producing an alternating current of approximately thirty milliamperes. It is to be noted that the inductance is remarkably constant for such large magnetomotive forces indicating a rather large air-gap. This reactor was designed to carry 300 milliamperes of direct current.

In a modified form of the circuit[1] the condenser is connected in series with the inductance to be measured, see Fig. 6, instead of in parallel as in the method already outlined. The measurement is made by throwing the single-pole switch first to left, noting the current through the inductance alone, and then to the right, noting the current through the capacity and inductance in series. If the line current is larger with the switch to the right the capacity is too great; if it is smaller the capacity is too small. The capacity is changed in accordance with these observations, and the switching operation repeated until there is no permanent change in the instrument reading. For a balance the capacity reactance is twice the inductive reactance and

$$L = \frac{1}{2w^2 C_{farad}} \text{ henries} \tag{8}$$

which is the same as given by (6). When measuring choke coils at 60 cycles and with C expressed in microfarads

$$L = \frac{3.52}{C} \text{ henries} \tag{9}$$

Also

$$C = \frac{3.52}{L}\ \mu\text{f} \tag{10}$$

From which it is seen that for $L = 100$ henries a capacity of 0.0352 μf is required, and for $L = 500$ henries there is required a capacity of 0.00704 μf. With the switch to the left the current lags by θ degrees, and with the switch to the right the current leads by the same amount.

In the first method, where measuring the inductance of two coils in parallel, the line current is the same as the current in the individual inductances (neglecting the slight effect of resistance), while in this case it is equal to twice the current through the individual inductances.

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[1] Mr. H. T. Lyman, Jr., graduate student in Communication Engineering, Yale University, used this circuit in obtaining data for the curves that follow.

Where it is necessary to use a line instrument of high resistance for measuring the alternating current, the second method is preferred. For measuring extremely small values of alternating current, either a crystal detector in series with a direct-current microammeter or an electron-tube voltmeter may be used across a resistance in the line.

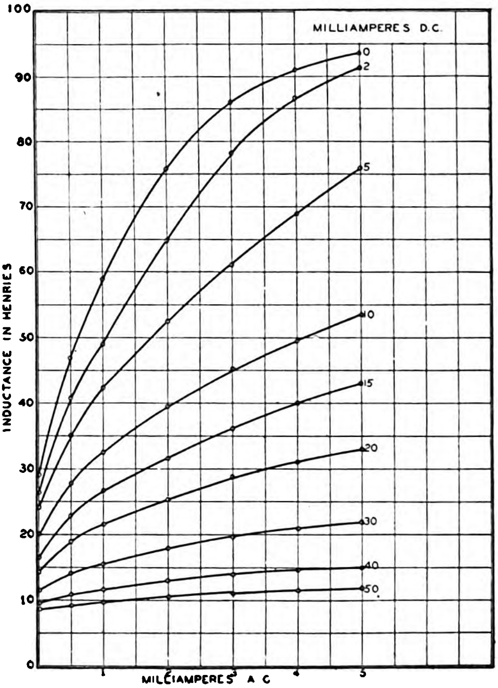

Fig. 7

The curves that follow show the variation of inductance with both the direct- and alternating-current magnetizing components through wide limits. The family in curves Figs. 6 and 7, which were plotted from the same data, have direct current as the independent variable in one case and alternating in the other. As a matter of convenience these are illustrated for a plate-filter-choke coil designed for a maximum full-load direct current of fifteen milliamperes; under normal operating conditions the alternating current will be small.

Fig. 8 is for the primary of an audio-frequency transformer used as a choke coil at 60 cycles.

These methods have been used to measure inductances of more than 3000 henries, and there is no reason why still larger values could not be measured if desired. They are simple, convenient to use, and require the minimum of equipment.

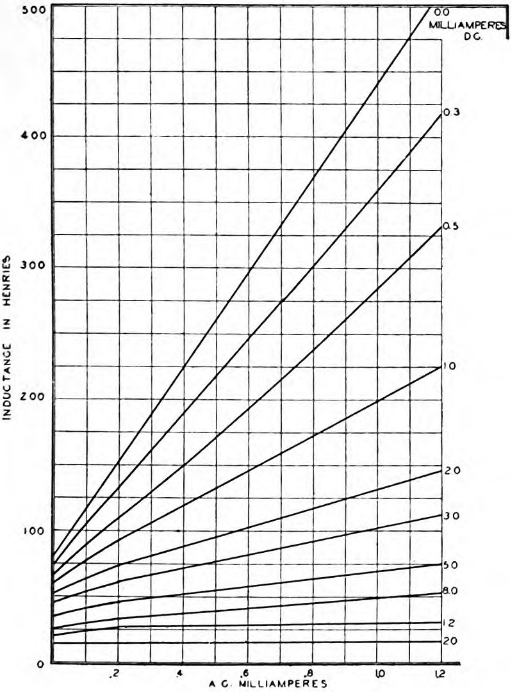

Fig. 8

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Volume 16, Number 11 *November, 1928*

FADING CURVES ALONG A MERIDIAN*

By

Robert C. Colwell

(Department of Physics, West Virginia University, Morgantown, West Virginia)

Summary—*The fluctuations in signal strength of KDKA, Pittsburgh, Pa., were observed through the sunset period of Morgantown, W. Va. Since the two cities are approximately on the same meridian, any variation at sunset should be due to changes in the Heaviside-Kennelly layer and not to refraction through the earth's shadow. Observations were made on twenty-one days; it was found that on bright clear days, the curve fluctuated considerably, while on cloudy days, the curve was fairly steady. The fluctuations can be accounted for by assuming a reflected wave on clear days.*

In May, 1925, measurements were made at Morgantown, West Virginia, upon the signal strength of Station KDKA at Pittsburgh, Pa. The results of these measurements were published by the Bureau of Standards.[1] The readings were continued during the winter of 1926–27, using a superheterodyne set connected to a semi-automatic recording galvanometer. These curves were averaged by the usual methods.

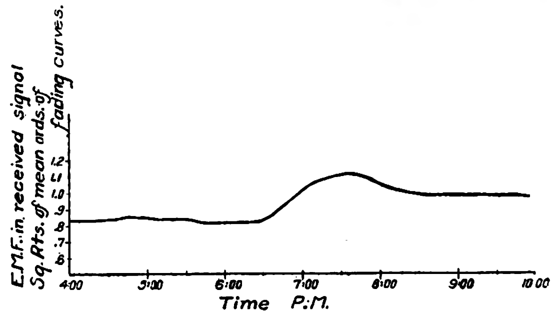

Fig. 1—Fading Curves, KDKA, West Virginia University, Morgantown, West Virginia.

All these curves were taken during the years 1926 and 1927 on the following dates:

Cloudy days 1926—January 10, 31; February 21.
1927—January 2, 9, 16, 30; February 20; March 7, 23, 24.
Clear days 1926—February 7, 23, 28; December 5.
1927—March 25, 26, 28, 29, 30, 31.

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* Original Manuscript Received by the Institute, July 11, 1928.

[1] Scientific Paper of the Bureau of Standards, No. 561, p. 432.

The curve for all the readings is given in Fig. 1. The time of day is plotted along the X axis, and it is apparent that there is a slight dip between the hours of six and seven which is near the sunset period. The curves were then averaged for five-minute intervals before and after sunset; that is, the time of sunset for each separate curve was taken as zero. This average for all the curves resulted in the curve shown in Fig. 2. This figure shows that the average strength remains fairly constant during the daylight hours, takes a slight dip shortly before sunset, and then rises to a night value about thirty per cent higher than the day value. The night value, however, is not constant and sometimes falls below that of the daylight hours.

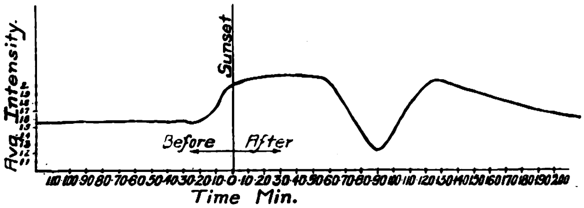

Fig. 2—Average Curve, KDKA, at Morgantown, West Virginia

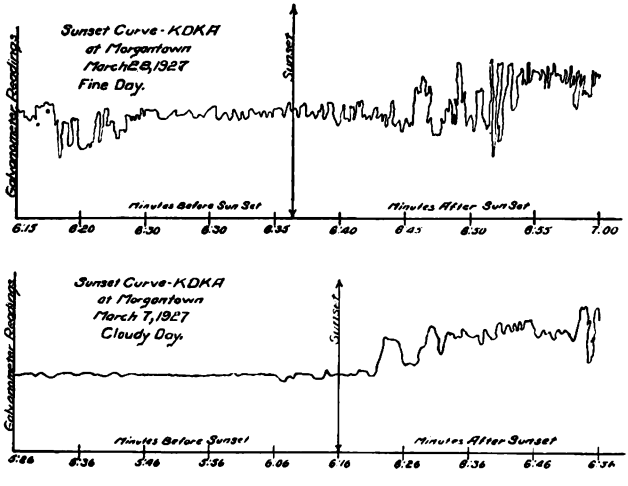

Fig. 3.

Upon examination of the curves, it was found that eleven curves had been taken on cloudy days and ten upon fine, clear days. The difference between the two types of curves is shown in

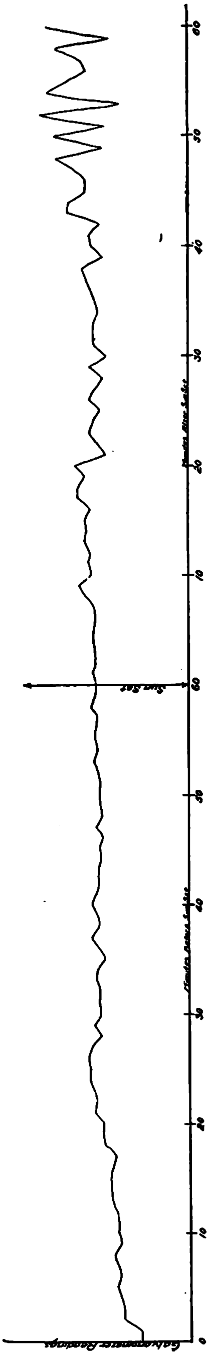

Fig. 4—Average Curve for Fine Days, KDKA, at Morgantown.

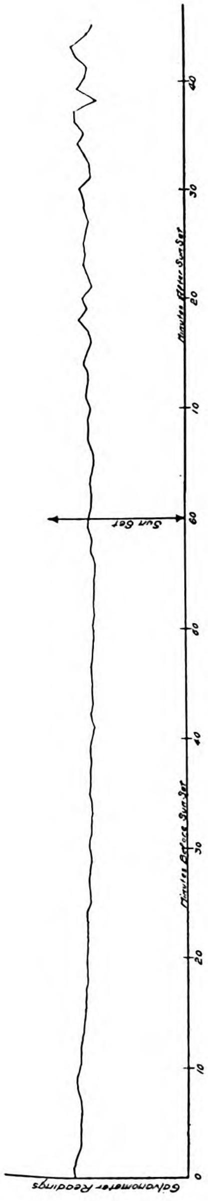

Fig. 5—Average Curve for Cloudy Days, KDKA, at Morgantown.

Fig. 3. It is apparent that the typical curve for a clear day is rather disturbed during the daylight hours, increases after sunset, and shows considerable fading. The typical curve for a cloudy day is uniform during the daylight hours, and the increase after night is fairly steady.

The ordinates for all the curves taken on fine, clear days were averaged at one-minute intervals before and after sunset. The curve of Fig. 4 shows the variable character of this average curve. A similar average was taken for the cloudy days and Fig. 5 shows the comparatively even curve typical of cloudy weather. These observations indicate a new relation between signal intensity and the state of the atmosphere. Other relations have been noted by Austin[2] and Pickard.[3]

The results of this investigation can be explained if it is assumed that the Kennelly-Heaviside layer is partially operative even during the daylight hours. On fine days there is a reflected wave (sky wave) which interferes with the ground wave causing a slight fluctuation during the afternoon. After sunset the reflected wave increases in intensity and fading becomes more pronounced. On cloudy days, the atmospheric conditions prevent the sky wave from reaching the reflecting layer, and only the steady ground wave is received.

It should be understood, however, that the typical cloudy weather curve can only be obtained in the middle of a cloudy period, and similarly for the fine weather curve. When the weather is changing from cloudy to clear and vice versa, the curves are very irregular and depart from the typical forms.

[2] Proc. I.R.E., 12, 681; December, 1924; 14, 781; December, 1926.
[3] Proc. I.R.E., 16, 765; June, 1928.

BOOK REVIEW

Theory of Vibrating Systems and Sound, BY IRVING B. CRANDALL. Published by D. Van Nostrand Company, New York, 1926. 272 pages, 23 illustrations, $6\frac{1}{4} \times 9\frac{1}{4}$ cloth, $5.00.

This book is based on studies in the Bell Telephone Laboratories and written by a well-known research worker in the field. It is intended to supplement, rather than to replace, the accepted treatises on sound by such men as Rayleigh, Lamb, and others. Its purpose is to present the theory of sound and its recent technical applications in such a way as to interest the student of physics who has given a certain amount of attention to analytical mechanics and to bring out the latest branch of applied science—"electro-acoustics."

The book also treats, in addition to problems of radiation and transmission, architectural acoustics and gives many references. The presentation is clear, as are all of Dr. Crandall's publications.

AUGUST HUND†

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RADIO STATIONS OF THE WORLD ON FREQUENCIES ABOVE 1500 KILOCYCLES*

THROUGH the courtesy of the Federal Radio Commission the following list of high-frequency allocations is published.

This list is published for the information of the Institute membership to indicate the present status of high-frequency assignments. It will not be published as a regular feature of the PROCEEDINGS.

The frequencies listed in the first column are those recently adopted by the Federal Radio Commission in its system of high-frequency channels. Each channel is approximately 0.2 per cent wide, and is indicated by the frequency at its center. Every frequency licensed or used by any station within each channel is given in the last column. This list shows only the channels upon which stations are operated. Intermediate channels, not shown, are separated by intervals similar to those shown, the separation being approximately two-tenths per cent.

The data for United States stations given correspond to the records of the Commission. For the stations of other nations, the data are not authoritative. They are compiled from various sources and have not been fully checked. In some cases, more than one frequency for a station is given because divergent values were given in the sources. In a few cases the frequencies are slightly uncertain because the source gave data in wavelengths, which may have been originally computed from frequencies by use of other conversion factors than 300,000.

For U. S. Government stations, only the frequency is given. This is because the Government departments have not released for publication the locations of their stations.

Since amateurs, and a few special cases of other stations, are licensed for bands of frequencies rather than specific channels, they are not mentioned in the table.

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
1500	C2BB	Valparaiso, Chile		1500
	VNAC	Port Elizabeth, U. of S. A.		1500
	CRLQ	Otchinjau, Angola		1500
	CRE	Dili, Portuguese Timor		1500
	C2AD	Valparaiso, Chile		1500
	C2BG	Vina del Mar, Chile		1500
	C3BK	Santiago, Chile		1500
	GKZ	Humber, England		1500
	——	Cuba (2 stations)		1500
1508	——	British Columbia, Canada (10 stations)		1510
1520	GMG	Guernsey, England		1520
1524	GMG	Guernsey, England		1523
	GCA	Tobermory, England		1523
	GCB	Lochboisdale, England		1523
1528	C1AC	Tocna, Chile		1530
1532	SMSM	Karlskrona, Sweden		1531

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* Original Manuscript Received by the Institute, October 11, 1928.

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
		Campbell River, B. C.		1540
1540	CO	Thurston Bay, B. C.		1540
	DM	Vina del Mar, Chile		1540
	C2BA	Swinemunde, Germany		1579
1580	DAS	Cuxhaven, Germany		1579
	DAC	Elbe Feuerschiff Vier, Germany		1579
	KBL	Valparaiso, Chile		1579
	C2AP	Valparaiso, Chile		1579
	C2AU	Antofagasta, Chile		1579
	C1AF	Lighthouse Service, Holland		1579
	—	Washington, D. C.	Jenkins Labs.	1605
1604	3XK	Providence, R. I.	C.E. Mfg. Co.	1605
	1XAC			
1612	—	Lighthouse Service, Holland		1613
1620	—			
1664	KGIA	Drummondville, P. Q.		1620
	KGIB	(Portable) Calif.	Geophysical Research Corp.	1664
	KGIC	(Portable) Calif.	Geophysical Research Corp.	1664
	KGID	(Portable) Calif.	Geophysical Research Corp.	1664
1668	SQJ	(Portable) Calif.	Geophysical Research Corp.	1664
	SQI	Amazonas, Brazil		1667
	C1AE	Amazonas, Brazil		1667
	C2BI	Antofagasta, Chile		1667
	C3CF	Vina del Mar, Chile		1667
1708	C3AA	Santiago, Chile		1667
	C3BJ	Santiago, Chile		1710
	GMF	Santiago, Chile		1710
1716	DAL	Alderney, England		1710
	POB2	Bremerhaven, Germany		1714
1720	PA25	Haarlem, Holland		1714
1760	C3BM	Medemblik, Holland		1720
	—	Santiago, Chile		1760
1764	C3AB	Tonning, Germany		1760
	C3AL	Santiago, Chile		1765
	C3BI	Santiago, Chile		1765
1788	C3CG	Santiago, Chile		1765
1808	PA24	Santiago, Chile		1790
1876	C3CB	Medemblik, Holland		1810
	BAC	Santiago, Chile		1875
	KBL	Cuxhaven, Germany		1875
1880	DAS	Elbe Feuerschiff, Vier, Ger.		1875
1900	PA23	Swinemunde, Germany		1875
1940	C3CC	Oude Zeng, Holland		1900
1976	3XQ	Santiago, Chile		1940
2000	CPO	Mt. Lakes, N. J.	S. Ballantine	1975
	CPN	Bolivia		2000
	—	Bolivia		2000
	PB4	Glace Bay, N. S.		2000
	PA22	Amsterdam, Holland		2000
2008	9BA	Breezand, Holland		2000
2020	—	Montreal, P. Q.		2010
2028	KVT	Manitoba, Canada (2 stations)		2020
2040	KLV	Los Angeles, Calif.	Boulevard Express	2030
	KLU	(Portable)	Lowell Observatory	2040
2048	KZI	Flagstaff, Arizona	Lowell Observatory	2040
	KFZ	(Portable) Calif.	Pratt & Dutro	2050
	KGV	(Portable) Calif.	Russell Reed	2050
	KFV	(Portable) Calif.	Russell Reed	2050
	KYY	(Portable) Calif.	L. A. Co. Forestry Dept.	2050
	KYX	Los Angeles, Calif.	L. A. Co. Forestry Dept.	2050
	KVP	(Portable) Calif.	Pratt & Dutro	2050
	—	Dallas, Texas	City of Dallas	2050
2052	KJA	Quebec, Canada (2 stations)		2050
2068	KVU	Pysht, Washington	Merrill & Ring Lbr. Co.	2054
	WCK	San Diego, Calif.	Boulevard Express	2070
	C3CE	Detroit, Mich.	Detroit Police Dept.	2070
2096	—	Santiago, Chile		2070
	3XI	Lightships		2097
2100	WEQ	Baltimore, Md.	G. E. Sterling	2098
	WJH	Baltimore, Md.	Brd. of Fire Comm.	2100
	WJX	Washington, D. C.	Potomac Elec. Pow. Co.	2100
	DU	Benning, D. C.	Potomac Elec. Pow. Co.	2100
	DJ	Drummondville, P. Q.		2100
	PA21	Yamachiche, P. Q.		2100
2108	WBV	Harlingen, Holland		2100
2140	2XBW	Charleroi, Pa.	W. Penn. Power Co.	2110
	3XE	Jersey City, N. J.	W. C. Von Brandt	2140
		Baltimore, Md.	Balto. Radio Show, Inc.	2140

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
2866	KGJ	Santa Barbara, Calif.	Merit, Chapman & Scott Cor.	2864
	KHR	Santa Barbara, Calif.	Merit, Chapman & Scott Cor.	2864
2884	——		U. S. Government	2885
2914	——		U. S. Government	2915
2938	9OB	Ottawa, Ont.		2940
2956	——		U. S. Government	2955
2962	——		U. S. Government	2960
	WSY	Darlington, Md.	Susquehanna Power Co.	2960
	——		U. S. Government	2965
2968	——		U. S. Government	2970
2974	——		U. S. Government	2975
2980	——		U. S. Government	2980
	DP	Provincial Air Base, Ont.		2980
	9AZ	Maple Mount, Ont.		2980
	9BD	Long Lake, Ont.		2980
	9BF	Timagami, Ont.		2980
	9BG	Savant Lake, Ont.		2980
	9BH	Cat Lake, Ont.		2980
	DO	Gold Pines, Ont.		2980
	DQ	Woman Lake, Ont.		2980
	DR	Red Lake, Ont.		2980
2992	——		U. S. Government	2995
2998	WEP	Cape Charles, Va.	Norfolk-Cape Charles Radio Telegraph Co.	3000
	9BH	Pine Ridge Post, Ont.		3000
	9BF	Sioux Lookout, Ont.		3000
	9BD	Red Lake, Ont.		3000
	9BG	Woman Lake, Ont.		3000
3004	——		U. S. Government	3005
3034	——		U. S. Government	3035
3052	WGI	Alpena, Mich.	Alpena Marine Radio Corp.	3050
3058	PB8	Rotterdam, Holland		3061
3064	——		U. S. Government	3065
3070	WKZ	Cumberland, Md.	Potomac Edison Co.	3070
3094	LA1M	Oslo, Norway		3093
	PB7	Groningen, Holland		3093
	——		U. S. Government	3095
3112	WJV	Philadelphia, Pa.	Philadelphia Elec. Co.	3110
3124	PB6	The Hague, Holland		3125
3154	——		U. S. Government	3155
3178	KHAH	*Greater Rockford* Monoplane	Bert Hassell	3178
3196	——		U. S. Government	3195
3202	DK	Shawinigan Falls, P. Q.		3200
	BG	Isle Naligme, P. Q.		3200
	BJ	Montreal, P. Q.		3200
	BH	Quebec, P. Q.		3200
	BI	St. Narcisse, P. Q.		3200
3208	3XK	Washington, D. C.	Jenkins Labs.	3210
	1XAC	Providence, R. I.	C.E. Mfg. Co.	3210
3226	——	Amsterdam, Holland		3226
3232	——		U. S. Government	3235
3262	——		U. S. Government	3265
3292	WFE	(Portable)	Cmdr. Richard E. Byrd	3290
	WFD	(Portable)	Cmdr. Richard E. Byrd	3290
	KFK	(Portable)	Cmdr. Richard E. Byrd	3290
	WFA	(Portable)	Cmdr. Richard E. Byrd	3290
	WFC	Airplane *Fairchild*	Cmdr. Richard E. Byrd	3290
	WFB	Airplane *Floyd Bennett*	Cmdr. Richard E. Byrd	3290
	WFF	Airplane *Fokker*	Cmdr. Richard E. Byrd	3290
	——		U. S. Government	3295
3298	PB1	Rotterdam, Holland		3297
	WBV	Charleroi, Pa.	West Penn. Power Co.	3301
	WOB	Connellsville, Pa.	West Penn. Power Co.	3301
3316	CM	Noranda, P. Q.		3320
	CK	Selquirk Mines, Manitoba		3320
3332	WPM	Birmingport, Ala.	Inland Waterways Corp.	3331
	——	Soerabaya, Java		3331
	——	Warsaw, Poland		3333
	——	Lwow, Poland		3333
3340	——		U. S. Government	3340
	9CH	Geizers Hill, Nova Scotia		3340
3348	——		U. S. Government	3345
	——		U. S. Government	3350
3356	——		U. S. Government	3355
	——		U. S. Government	3360
3364	——		U. S. Government	3365
3372	——		U. S. Government	3370
	——		U. S. Government	3375

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
3380	——		U. S. Government	3380
3388	——		U. S. Government	3385
	——		U. S. Government	3390
3396	——		U. S. Government	3395
	——		U. S. Government	3400
3404	——		U. S. Government	3405
3412	——		U. S. Government	3410
	——		U. S. Government	3415
3428	WHF	Williamsport, Md.	Potomac Edison Co.	3430
3444	——		U. S. Government	3445
3468	KTT	Drier Bay, Alaska	Gorman Packing Corp.	3466
3476	——		U. S. Government	3475
	9BG	Savant Lake, Ont.		3480
	9BD	Long Lake, Ont.		3480
	9BH	Cat Lake, Ont.		3480
	9AZ	Maple Mount, Ont.		3480
	9BF	Timagami, Ont.		3480
3500	——		U. S. Government	3500
3532	LA1E	Bergen, Norway		3529
	C2BD	Valparaiso, Chile		3530
3588	RZV	Vonjama, Liberia		3590
3612	RZS	Sinoe, Liberia		3610
3660	RZB	Bassa, Liberia		3660
3700	RZP	Cape Palmas, Liberia		3700
3716	HBC	Berne, Switzerland		3720
3748	KFUH	SS *Kaimiloa*	M. R. Kellum	3748
	KGDQ	SS *Faith*	W. W. Shaw	3748
	HVA1	Hanoi, Indo-China		3750
	RZM	Monrovia, Liberia		3750
	HJF	Cucuta, Columbia (Proposed)		3750
	HJE	Cali, Columbia (Proposed)		3750
	VMDJ	White Island, New Zealand		3750
	2FF	Tientsin, China		3750
	XWAG	Vancouver, B. C.		3750
	——	Calgary, Alberta		3750
	——	Saskatoon, Sask.		3750
	XWAC	Winnipeg, Manitoba		3750
	——	London, Ont.		3750
	——	Toronto, Ont.		3750
	XWCB	Camp Borden, Ont.		3750
	——	Kingston, Ont.		3750
	——	Montreal, P. Q.		3750
	——	Quebec, P. Q.		3750
	——	St. John, N. B.		3750
	——	Halifax, N. S.		3750
3900	KHAS	Airplane *Maid of Detroit*	H. C. McCarroll	3904
3948	GSA			3945
	3XQ	Mt. Lakes, N. J.	S. Ballantine	3950
	JCX	Naha, Japan		3950
3956	——	Paris, France		3953
3996	YZ	Issy-les-Moulineaux, France		3998
	FL	Paris, France		3998
	JKV	Kanazawa, Japan		4000
	SQBP	Pernambuco, Brazil		4000
	SQBQ	Pernambuco, Brazil		4000
4012	WLC	Rogers, Mich.	Michigan Limestone Co.	4010
	——		U. S. Government	4015
4020	——		U. S. Government	4017
	——		U. S. Government	4020
4028	——		U. S. Government	4025
	——		U. S. Government	4030
4044	KOG	Honolulu, Hawaii	Mutual Telephone Co.	4044
	——		U. S. Government	4045
4052	——		U. S. Government	4050
	KFVM	SS *Idalia*	E. R. Parker	4052
	——		U. S. Government	4055
	JHL	Hiroshima, Japan		4050
4060	——		U. S. Government	4060
4068	——		U. S. Government	4065
	——		U. S. Government	4070
4076	——		U. S. Government	4075
	——		U. S. Government	4080
	JBD	Keijo, Japan		4080
4084	——		U. S. Government	4085
4092	——		U. S. Government	4090
4100	8XA	Any Ford plane	Ford Motor Co.	4100
	IDM	Rhodes, Italy		4100

Channel	Call Letters	Location of Transmitter	Owner	Frequency
4100	JYD	Tokyo, Japan		4100
4108	—		U. S. Government	4105
	—	Yucatan		4110
	JPP	Tokyo, Japan		4110
4132	—		U. S. Government	4135
4140	KHM	Lihue, Hawaii	Mutual Telephone Co.	4144
	KHO	Kaunakakai, Hawaii	Mutual Telephone Co.	4144
	KHL	Wailuku, Hawaii	Mutual Telephone Co.	4144
	KLN	Hilo, Hawaii	Mutual Telephone Co.	4144
4148	JKF	Kuji, Japan (Proposed)		4150
4156	—		U. S. Government	4155
4204	—		U. S. Government	4205
4220	KHAI	Airplane *C-5358*	Western Air Express	4224
	KHAG	Airplane *C-5177*	Western Air Express	4224
	KHAJ	Airplane *C-4458*	Western Air Express	4224
	KGHY	Mills Field, San Francisco, Calif.	Western Air Express	4224
4228	JEW	Osaka, Japan		4230
4236	—		U. S. Government	4235
4252	WRB	Miami, Fla.	W. G. Watts, Jr.	4250
	WRP	Pinecrest, Fla.	W. G. Watts, Jr.	4250
	—		U. S. Government	4255
4268	—		U. S. Government	4265
4276	1XAC	Providence, R. I.	CE Mfg. Co.	4280
	6XT	San Francisco, Calif.	C. L. Watson & R. C. Gray	4280
	2XQ	New York, N. Y.	Univ. Wireless Commun. Co.	4280
	9XC	Chicago, Ill.	Univ. Wireless Commun. Co.	4280
	2XDY	Jersey City, N. J.	Walter C. Von Brandt	4280
	3XE	Baltimore, Md.	Baltimore Radio Show	4280
	IDZ	Rome, Italy		4280
4284	—	Cadiz, Spain		4286
	—	Matagorda, Spain		4286
4292	SQAT	Campina Grande, Parahyba, Brazil		4290
	JBK	Kogoshima, Japan		4290
	SQAS	Rio Branco, Pernambuco, Brazil		4290
	SQAU	Parahyba, Brazil		4290
	SQAR	Recife, Pernambuco, Brazil		4290
	SQBS	Limoeiro, Pernambuco, Brazil		4290
	—		U. S. Government	4295
4300	—		U. S. Government	4300
4308	—		U. S. Government	4305
4316	—		U. S. Government	4310
4348	JMPA	Saishu, Japan		4350
4364	—		U. S. Government	4365
4372	—		U. S. Government	4370
	—		U. S. Government	4375
4380	—		U. S. Government	4380
4388	—		U. S. Government	4385
4396	KHAS	Airplane *Maid of Detroit*	H. G. McCarroll	4400
4405	WFB	Airplane *Floyd Bennett*	Cmdr. Richard E. Byrd	4405
	WFF	Airplane *Fokker*	Cmdr. Richard E. Byrd	4405
	WFC	Airplane *Fairchild*	Cmdr. Richard E. Byrd	4405
	KFK	(Portable)	Cmdr. Richard E. Byrd	4405
	WFA	(Portable)	Cmdr. Richard E. Byrd	4405
	WFE	(Portable)	Cmdr. Richard E. Byrd	4405
	WFD	(Portable)	Cmdr. Richard E. Byrd	4405
4425	—		U. S. Government	4430
4435	—		U. S. Government	4435
	—		U. S. Government	4436
	—		U. S. Government	4440
4445	—		U. S. Government	4445
	JMCA	Mokpo, Japan		4450
4505	KWV	Bakers Field, Calif.	Pacific Air Transport	4510
4515	KTF	Midway Island	Mackay Radio & Teleg. Co.	4515
4525	—		U. S. Government	4525
	9AQ	Nipigon, Ont.		4530
	9AI	Toronto, Ont.		4530
4535	6XAI	Los Angeles, Calif.	Los Angeles Radio Club	4540
	8XF	Cleveland, Ohio	Radio Air Service Corp.	4540
4545	RZM	Monrovia, Liberia		4550
4565	RZP	Cape Palmas, Liberia		4570
4575	9XAD	Chicago, Ill.	J. G. Branch	4575
4595	WGT	San Juan, P. R.	Radio Corp. of America	4593
	2XBA	Newark, N. J.	WAAM Inc.	4600

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
4595	——	Halifax, N. S.		4600
	——	Charlottetown, P. E. I.		4600
	——	Sherbrooke, P. Q.		4600
	——	Ottawa, Ont.		4600
	XWCB	Camp Borden, Canada		4600
	——	Hamilton, Ont.		4600
	——	Chatham, Ont.		4600
	——	Port Arthur, Ont.		4600
	——	Regina, Sask.		4600
	XWBC	Edmonton, Alta.		4600
	——	Victoria, B. C.		4600
	——	Boko, Japan (Proposed)		4600
	——	Tainan, Japan (Proposed)		4600
4685	2XE	Richmond Hill, N. Y.	Atlantic Brdcstg. Corp.	4685
	SQAP	Belem, Brazil		4690
	SQAQ	Arumanduba, Brazil		4690
4695	1XAB	Portland, Me.	H. P. Rines	4700
	——	Boko, Japan (Proposed)		4700
	——	Tainan, Japan (Proposed)		4700
	DF	Winnipeg, Man.		4700
	DE	Hudson, Ont.		4700
	DP	The Pas, Man.		4700
	DA	Cold Lake, Man.		4700
	UL	Managua, Nicaragua		4700
4755	SQBI	Recife, Pernambuco, Brazil		4760
4765	SQBT	Serra Grande Alagoas, Brazil		4760
4835	G5DH	Dollis Hill, England		4835
	——	Taihoku, Japan (Proposed)		4840
4895	JEW	Osaka, Japan		4900
	UQ	Bluefields, Nicaragua		4900
4905	9XU	Council Bluffs, Iowa	Mona Motor Oil Co.	4910
4915	CF	Drummondville, P. Q.		4916
	JEW	Osaka, Japan		4920
4935	CG	Drummondville, P. Q.		4937
4975	AGC	Nauen, Germany		4975
4985	UOG	Vienna, Austria		4983
	IDX	Asmara, Italy		4990
	ISL	Afgoi, Italian Somaliland		4990
4995	CF	Drummondville, P. Q.		4997
	3XL	Bound Brook, N. J.	Radio Corp. of America	5000
	——	Warsaw, Poland		5000
	——	Lwow, Poland		5000
	JPS	Sapporo, Japan		5000
5045	JFBB	Giran, Japan		5050
5055	——	Spain		5060
5065	JBD	Keijo, Japan		5070
5095	WKA	East Pittsburgh, Pa.	Westinghouse Elec. & Mfg. Co.	5100
	UW	Cape Gracias, Nicaragua		5100
5115	JCX	Naha, Japan (Proposed)		5120
5135	——	Spain		5140
5155	KWT	Palo Alto, Calif.	Mackay Radio & Teleg. Co.	5160
5165	JHL	Hiroshima, Japan		5170
	——	Nauen, Germany		5170
5245	JPP	Tokyo, Japan		5250
	——	Managua, Nicaragua	Emergency & Special Service	5250
5255	XMT	Mobile on Railroad, China		5260
5265	JPP	Tokyo, Japan		5270
5295	——	Chosi, Japan (Proposed)		5300
	——	Managua, Nicaragua	Emergency & Special Service	5300
5345	JKV	Kanazawa, Japan		5350
5355	GBM	Leafield, England		5355
	——	Oxford, England		5360
	JKV	Kanazawa, Japan		5360
5395	JKF	Kuji, Japan (Proposed)		5400
5405	——	Chigiqui, Panama	Tropical Radio Teleg. Co.	5410
5445	——	Riga, Latvia		5450
	IDO	Rome, Italy		5450
	JRTS	Tientsin, China		5450
5455	JEW	Osaka, Japan		5460
5485	CA	Geizers Hill, N. S.		5490
5495	JYZ	Tokyo, Japan		5500
	——	SS *Fort James*, Canada		5500
5525	WRL	Duluth, Minn.	Radio Corp. of America	5525
	WCY	Cleveland, Ohio	Radio Corp. of America	5525
	WGO	South Chicago, Ill.	Ill. Radio Corp. of America	5525
5540	KRK	Palo Alto, Calif.	Mackay Radio & Teleg. Co.	5533
5555	7XO	Seattle, Wash.	Northwest Radio Service Co.	5550
	2XBH	Brooklyn, N. Y.	Charles G. Unger	5550

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Fre-quency*
7505	LVF	Flores, Guatemala		7500
	FSQ	Sainte Assise, France		7500
	B82	Belgium		7500
	ISV	Italian Somaliland		7500
	OCDB	Abyssinia		7500
	SQCL	Brazil		7500
	PB9	Holland		7500
	7MN	Cuba		7500
	——	Mexico		7500
	JFBB	Japan		7500
	BAM	Tahiti		7500
7520	WEG	Rocky Point, N. Y.	Radio Corp. of America	7520
7535	DIY	Königs Wüsterhausen, Germany		7537
7565	——	Chiriqui, Panama	Tropical Radio Teleg. Co.	7560
	AFU	Königs Wüsterhausen, Germany		7560
	RZA	Monrovia, Liberia		7570
7595	YR	La Dona, France		7595
	JFBB	Giran, Japan		7600
	UL	Managua, Nicaragua		7600
	OCLY	France		7600
7610	——	Nauen, Germany		7610
7625	——	Milwaukee, Wis.	American Publishers Comm.	7625
	——	Los Angeles, Calif.	American Publishers Comm.	7625
7640	——	Washington, D. C.	American Publishers Comm.	7640
	——	New York, N. Y.	American Publishers Comm.	7640
	OCDB	Jibuti, Fr. Somaliland		7633
7670	——	Sayville, N. Y.	Mackay Radio & Teleg. Co.	7670
7685	4JE	San Juan, P. R.		7688
	——	Nauen, Germany		7690
	JPP	Tokyo, Japan		7690
	SQCL	Sao Paulo, Brazil		7690
	OCDB	Abyssinia		7690
	IDX	Italy		7690
	——	Mexico		7690
	XOM	China		7690
	PWA	Cuba		7690
	——	Denmark		7690
	OCMV	France		7690
	OCRU	Africa		7690
	HZU	Port Gentil		7692
7700	JPP	Tokyo, Japan		7700
	JYB	Tokyo, Japan		7700
	UQ	Bluefields, Nicaragua		7700
7715	KEE	Bolinas, Calif.	Radio Corp. of America	7715
	BYA	Whitehall, England		7720
7730	——	Sayville, N. Y.	Mackay Radio & Teleg. Co.	7730
	——	Kootwijk, Holland		7732
7745	KNW	Palo Alto, Calif.	Mackay Radio & Teleg. Co.	7745
7760	KNN	Honolulu, Hawaii	Mackay Radio & Teleg. Co.	7760
	——	Kootwijk, Holland		7765
7775	WMU	Elizabeth, N. J.	Standard Oil Co. of N. J.	7775
	WTF	Akron, Ohio	Firestone Plantations Co.	7775
7790	B82	Brussels, Belgium		7790
	XDA	Mexico		7790
	AND	Java		7790
	FUE	France		7790
	OXZ	Denmark		7790
	——	Buenos Aires, Argentine		7790
7805	JRV	Saipan, Japan		7800
	——	Nauen, Germany		7800
	UW	Cape Gracias, Nicaragua		7800
	FW	France		7800
7820	——	Chicago, Ill.	American Publishers Comm.	7820
	DFT	Nauen, Germany		7820
7835	——	New York, N. Y.	American Publishers Comm.	7835
	——	Kootwijk, Holland		7835
7850	——	Floral Park, N. Y.	American Publishers Comm.	7850
7865	SUX	Abu Zabal, Egypt (Proposed)		7860
7880	——	Drummondville, P. Q.		7880
7895	KFUH	SS *Kaimiloa*	M. R. Kellum	7890
	C3AN	Santiago, Chile		7890
	JPS	Sapporo, Japan		7890
	ISL	Afgoi, Italian Somaliland		7890
	C2BK	Valparaiso, Chile		7890
	SQK	Maraba, Para, Brazil		7890
	SQY	Conceigao de Araguaya, Brazil		7890
	SQZ	Soure, Brazil		7890
	CF	Drummondville, P. Q.		7890

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
7895	CG	Drummondville, P. Q.		7890
	CJ	Drummondville, P. Q.		7890
	—	Mexicali		7890
	VQF	Sarawac		7890
	PKH	Java		7890
7895	IST	Afgoi, Italian Somaliland		7895
	ISQ	Afgoi, Italian Somaliland		7895
	3XQ	Mt. Lakes, N. J.	Stuart Ballantine	7900
	HBC	Berne, Switzerland		7900
	—	Nauen, Germany		7900
	JPS	Sapporo, Japan		7900
	—	Emergency & Spec. Service Nic.		7900
7910	DFP	Nauen, Germany		7910
7925	—	New York, N. Y.	American Publishers Comm.	7925
7940	GLM	Dorchester, England		7940
7955	—	Chicago, Ill.	American Publishers Comm.	7955
	DGL	Nauen, Germany		7960
7985	—	Nauen, Germany		7980
	FL	France		7980
8000	KFZQ	SS *Robador*, USA	Robert Law, Jr.	7995
	AND	Java		7995
	6XAS	Calif. (Portable)	Culver City Ra. Elec. Co.	8000
	C1AH	Chuquicamata, Chile		8000
	C3AD	Santiago, Chile		8000
	C2BJ	Valparaiso, Chile		8000
	C2BM	Quilpue, Chile		8000
	JKV	Kanazawa, Japan		8000
	JBR	Siberia, Maru, Japan		8000
	XKY	Kwaiyang, China		8000
	XRO	Chungking, China		8000
	C3CK	Rancagua, Chile		8000
	HBC	Berne, Switzerland		8000
	SDA	San Salvador		8000
	—	Mexico		8000
	ANF	Java		8000
	XOM	China		8000
	GLK	Dorchester, England		8005
8015	WLC	Rogers, Mich.	Mich. Limestone Co.	8010
8030	—		U. S. Government	8030
	—		U. S. Government	8034
8045	—		U. S. Government	8040
	—		U. S. Government	8050
	WCFL	Chicago, Ill.	Chicago Fed. of Labor	8050
	HJG	Bogota, Colombia		8050
8060	—		U. S. Government	8060
8075	KNN	Honolulu, Hawaii	Mackay Radio & Teleg. Co.	8075
8090	—		U. S. Government	8090
8105	6XF	Los Angeles, Calif.	Calvin J. Smith	8100
	WJD	New York, N. Y.	International News Service	8100
	KUY	Bear Creek, Alaska	New York-Alaska Gold Dredging Co.	8100
	—		U. S. Government	8100
	KGBT	SS *Enchantress*	S. F. Wainwright	8103
	KFVM	SS *Idalia*	E. R. Parker	8103
	KFHW	SS *Poinsettia*	Walter W. Horne	8103
	KGBB	SS *Sachen*	R. B. Metcalf	8103
	—		U. S. Government	8110
	GKT	Bunham, England		8110
	C2BL	Valparaiso, Chile		8110
	C1AI	Chuquicamata, Chile		8110
	KAV	Norddeich, Germany		8110
8120	—		U. S. Government	8120
	—	Manila, P. I.		8120
8135	—		U. S. Government	8135
	—		U. S. Government	8140
8150	—		U. S. Government	8150
	UR	Cartago, Costa Rica	Tropical Radio Teleg. Co.	8150
	OIC	Ship, Austria		8150
8165	—		U. S. Government	8160
	—		U. S. Government	8170
	DGV	Nauen, Germany		8170
8180	—	Kootwijk, Holland		8174
	DIX	Königs Wüsterhausen, Germany		8179
	—		U. S. Government	8180
8195	VDB	Esquimalt, Canada		8190
8210	KGH	Hillsboro, Oregon	Mackay Radio & Teleg. Co.	8210
	—		U. S. Government	8210

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
8210	GKT	Burnham, England		8210
	ME5	Matagalpa, Nic. (Proposed)		8210
	ME7	Ocotal, Nic. (Proposed)		8210
	BLF	Bluefields, Nicaragua		8210
	CAB	Puerto Cabezas, Nicaragua		8210
	F8M	(Aviation) Nicaragua		8210
	GLKY	SS. *Carinthia*, England		8214
	C3BG	Santiago, Chile		8220
	—	Mexico		8220
	ANF	Java		8220
	FUT	France		8220
	SAV	Sweden		8220
8230	GKS	Dollis Hill, England		8230
	HIK	Barahona, San Domingo		8237
	WJD	New York, N. Y.	New York Times Co.	8230
8250	SAB	Goteborg, Sweden		8242
8270	—		U. S. Government	8270
	IR1	Italy		8275
	SPC	Brazil		8280
8290	—	Königs Wüsterhausen, Germany		8300
	HBC	Berne, Switzerland		8300
	UL	Managua, Nicaragua		8300
8310	—		U. S. Government	8310
	IDX	Asmara, Italy		8320
8330	KTA	Guam, Mariana Island	Mackay Radio & Teleg. Co.	8328
	DAN	Norddeich, Germany		8330
	KCR	Riga, Latvia		8330
	FQ	Leopoldville, Congo, Belgium		8330
	C2BO	Vina del Mar, Chile		8330
	OCRB	Morocco		8330
	PCMM	Holland		8330
	XVX	Papeete, Tahiti, French Oceania		8333
8370	WME	Duluth, Minn.	Intercity Radio Teleg. Co.	8370
	WDI	Detroit, Mich.	Intercity Radio Teleg. Co.	8370
	WAM	Buffalo, N. Y.	Intercity Radio Teleg. Co.	8370
	WTL	Cleveland, Ohio	Intercity Radio Teleg. Co.	8370
8390	WRL	Duluth, Minn.	Radio Corp. of America	8385
	WGO	South Chicago, Ill.	Ill. Corporation of America	8385
	WCY	Cleveland, Ohio	Radio Corp. of America	8385
	—	Königs Wüsterhausen, Germany		8400
	UQ	Bluefields, Nicaragua		8400
8410	—		U. S. Government	8410
8430	FUT	Toulon, France		8422
	WSL	Sayville, N. Y.	Mackay Radio & Teleg. Co.	8425
	SNNI	Rio de Janeiro		8437
8450	PCRR	Sgravenhage, Holland		8446
	UR	Costa Rica		8450
	BZC	England		8450
	RKU	Russia		8450
	ICF	Italy		8450
8470	—		U. S. Government	8470
	—	Nauen, Germany		8480
8490	G5DH	England		8495
	PCH	Scheveningen, Holland		8499
	UW	Cape Gracias, Nicaragua		8500
	GCA	England		8500
8510	—		U. S. Government	8510
	RXY	SS *City of San Francisco*		8518
8530	—		U. S. Government	8530
8550	SQI	Brazil		8540
	OCKO	Cotonou, Dahoney, Fr. W. Africa		8550
8570	KFUH	SS *Kaimiloa*, USA	M. R. Kellum	8566
	BZE	Matura Ceylon	British Naval Station	8566
	BYZ	Rinella, Malta	British Naval Station	8566
	CH	Santiago, Chile		8566
	OCRB	Morocco		8566
	ASD	SS *Bergsdalen*, Norway		8566
	BZF	Aden Somaliland	British Naval Station	8566
	BXW	Seletar, Singapore	British Naval Station	8566
	CKN	Esquimalt, B. C.		8570
	C3AU	Santiago, Chile		8570
	BXW	Seletar Straits Settlements		8570
	C2AH	Valparaiso, Chile		8570
	2AA	Tientsin, China		8570
	RAU	Russia		8570
	OCDA	French West Africa		8570

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
8570	IDB	Albania		8570
	PCA	Holland		8570
	2RM	Cuba		8570
	2JT	Cuba		8570
	UR	Costa Rica		8570
	G5DH	Dollis Hill, England		8570
	——	Mexico		8570
	——	Yucatan		8570
	1FMH	San Salvador		8571
	BYC	Horsea, England		8571
	——	Tamatave, France		8571
	BXY	Hongkong, China		8571
	BWW	Gibraltar		8571
	BYB	Whitehall, England		8571
	C3CJ	Santiago, Chile		8580
8590	——		U. S. Government	8590
	——		U. S. Government	8600
	KWT	Palo Alto, Calif.	Mackay Radio & Teleg. Co.	8600
8610	——		U. S. Government	8610
	——		U. S. Government	8620
	CTO	Portugal		8620
	CKN	British Protectorate		8620
8630	BYB	Cleethorpes, England		8630
8650	9XC	Chicago, Ill.	Univ. Wireless Comm. Co.	8650
	2XQ	New York, N. Y.	Univ. Wireless Comm. Co.	8650
	3XE	Baltimore, Md.	Balto. Radio Show, Inc.	8650
	6XT	San Francisco, Calif.	C. L. Watson & R. C. Gray	8650
	6XD	Portable	D. B. McGown	8650
	1XAC	Providence, R. I.	C. E. Manufacturing Co.	8650
	2XBY	Jersey City, N. J.	Walter C. Von Brandt	8650
	LPX	Monte Grande, Argentine		8645
8670	DIW	Königs Wüsterhausen, Germany		8670
8690	PJC	Curacao, Dutch West Indies		8690
	GLKY	SS *Carinthia*, England		8690
	LPI	Buenos Aires, Argentine		8690
	C2AK	Vina del Mar, Chile		8690
	C3CL	Santiago, Chile		8700
	HBC	Berne, Switzerland		8700
	HBC	Münchenbuchsee, Switzerland		8700
	VWZ	India		8700
	AYF	Venezuela		8700
	VIS	Australia		8700
8710	KNN	Honolulu, Hawaii	Mackay Radio & Teleg. Co.	8720
8730	——		U. S. Government	8730
	——		U. S. Government	8740
8750	——		U. S. Government	8750
	LCHO	Oslo, Norway		8750
	OCX	Morocco		8750
	RZA	Monrovia, Liberia		8750
	——		U. S. Government	8760
8770	——		U. S. Government	8770
	BXC	Nore, England		8770
	GBI	Grimsby, England		8780
8790	CF	Drummondville, P. Q.		8795
	CFH	Halifax, Nova Scotia		8795
	KHAS	Airplane *Maid of Detroit*, USA	H. G. McCarroll	8800
	——	India		8800
	——	England		8800
8810	WFC	Airplane *Fairchild*	Cmdr. Richard E. Byrd	8810
	WFF	Airplane *Fokker*	Cmdr. Richard E. Byrd	8810
	WFB	Airplane *Floyd Bennett*	Cmdr. Richard E. Byrd	8810
	WFA	(Portable)	Cmdr. Richard E. Byrd	8810
	WFE	(Portable)	Cmdr. Richard E. Byrd	8810
	WFD	(Portable)	Cmdr. Richard E. Byrd	8810
	KFK	(Portable)	Cmdr. Richard E. Byrd	8810
	IDM	Rhodes, Italy		8810
	IDO	Rome, Italy		8810
	PCA	Amsterdam, Holland		8818
	LGN	Norway		8818
	GBJ	Bodmin, England		8820
	CRA	San Jose, Costa Rica		8820
	SLO	Stanleyville, Congo, Belgium		8820
	PCUU	The Hague, Holland		8820
	QCK	Tientsin, China		8820
	——	SS *Jervis Bay*		8820
	XDA	Mexico		8820
	RKV	Russia		8820
	VPQ	British East Indies		8820

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Fre-quenc*
9230	OCDJ	Issy les Moulineaux, France		9230
	PCTT	Kootwijk, Holland		9240
	——	Kootwijk, Holland		9240
9250	GBK	Bodmin, England		9260
9270	——	Spain		9280
9290	——	Naha, Japan (Proposed)		9300
	UW	Cape Gracias, Nicaragua		9300
	AYG	Venezuela		9300
9310	AYA	Venezuela		9320
9330	CG	Drummondville, P. Q.		9332
	CG	Quebec, P. Q.		9338
9350	JCX	Naha, Japan		9360
9370	BIH	Paris, France		9369
	PTT	Brasil		9369
	FL	Paris, France		9369
	VIS	Sydney, Australia		9369
	ARDI	SS *C. A. Larsen*		9369
	8XAO	Detroit, Mich.	Station WJR Inc.	9370
	ANE	Java		9370
	JHL	Japan		9370
	VJC	Raboul, New Guinea		9375
	2MK	Cuba		9380
	5CX	Cuba		9380
	HVA	Tonkin, French Indo-China		9380
	H9OC	Switzerland		9380
	LY	France		9380
	OCNG	France		9380
	VQF	Sarawak		9380
	PWA	Cuba		9380
	——	England		9380
	ANH	Java		9380
	H9XD	Switzerland		9380
	JB	Africa		9380
	PCLL	Holland		9380
	PKD	Pokong, Dutch W. I.		9380
	PKX	Java		9380
	OCDJ	France		9375
	IDO	Rome, Italy		9370
	LDW	Bodoe, Norway		9375
	XVX	Papeete, Tahiti, French Oceania		9375
	OCGB	Ivory Coast, French West Africa		9375
	JHL	Hiroshima, Japan		9380
	——	Norway (Proposed)		9380
9390	——	Königs Wüsterhausen, Germany		9400
	——	Nicaragua (Emergency)		9400
9410	WGA	New York City	The Robert Dollar Co.	9410
	KGX	Los Angeles, Calif.	The Robert Dollar Co.	9410
	KGS	Honolulu, Hawaii	The Robert Dollar Co.	9410
	KGR	Seattle, Wash.	The Robert Dollar Co.	9410
	KGQ	San Francisco, Calif.	The Robert Dollar Co.	9410
9430	WEIR	Rocky Point, N. Y.	Radio Corp. of America	9430
	SAS	Carlsborg, Sweden		9434
	——	Nauen, Germany		9440
9450	WES	Rocky Point, N. Y.	Radio Corp. of America	9450
	——	Horomushiro, Japan (Proposed)		9450
	DIU	Königs Wüsterhausen, Germany		9458
9470	WET	Rocky Point, N. Y.	Radio Corp. of America	9470
9490	PCPP	Kootwijk, Holland		9488
	WEF	Rocky Point, N. Y.	Radio Corp. of America	9490
	——	Taihoku, Japan (Proposed)		9500
9510	PTR	SS *N. T. Nielsen Alonso*, Norway		9518
	ARCX	SS *Nilson Alonzo*		9518
	——	Australia		9520
9530	BVJ	England		9525
	PTK	Brazil		9525
	PKP	Java		9525
	SNM	Sweden		9525
9550	UL	Managua, Nicaragua		9550
	——	Brazil		9555
9570	——	Königs Wüsterhausen, Germany		9580
	——	Norway		9580
9590	LGN	Oslo, Norway		9600
	FW	France		9600
9610	LCK	Oslo, Norway		9620
	LDW	Norway		9620

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
9630	CF	Drummondville, P. Q.		9630
	CG	Drummondville, P. Q.		9630
	CJ	Drummondville, P. Q.		9630
9650	OCRU	West Africa		9645
	DGU	Nauen, Germany		9650
	2KP	Cuba		9650
	8XAG	Dayton, Ohio	E. T. Flewelling	9660
9670	JBD	Keijo, Japan		9670
	SAD	Sweden		9670
	TUE	SS *William Blumer*, Holland		9672
	SDA	SS *Ragunda*, Sweden		9672
	AGC	Nauen, Germany		9672
	——	Tananarive, France		9677
	AQS	Jan Mayen, Norway		9678
9690	DFF	Nauen, Germany		9690
	2XAL	Coytesville, N. J.	Experimenter Publishing Co.	9700
	——	Königs Wüsterhausen, Germany		9700
	——	Argentine		9700
	ELFP	Monrovia, Liberia		9700
9710	VLB	New Zealand		9710
	PTT	Brazil		9710
9730	DFO	Nauen, Germany		9730
	XDA	Mexico		9725
9750	WNC	Ocean Township, N. J.	American Tel .& Tel. Co.	9750
9770	DFS	Nauen, Germany		9770
	RKI	Russia		9770
	EAM	Spain		9770
9790	——	Argentine		9790
	GBW	Rugby, England		9790
	JKF	Kuji, Japan (Proposed)		9800
9810	DFE	Nauen, Germany		9810
	RCI	Russia		9810
9830	PTQ	Brazil		9835
	LCM	Norway		9835
	——	Nauen, Germany		9840
9850	PJC	Curacao, Dutch West Indies		9850
9870	WMI	Deal, N. J.	American Tel. & Tel. Co.	9870
9890	DGC	Nauen, Germany		9890
	HZAI	Saigon		9900
	SQBD	Brazil		9900
	PWAL	Brazil		9900
	VMF	New Zealand		9950
9910	EAX	Spain		9910
9930	——	Nauen, Germany		9930
	——	Spain		9940
	PCJJ	Holland		9930
	——	France		9930
	PJD	St. Martins		9940
9950	GBU	Rugby, England		9950
9970	KZET	Manila, P. I.	Radio Corp. of Philippines	9961
	IPR	Italy		9975
9990	GBM	Oxford, England		9994
	FTJ	Jacques Cartier, France		9994
	PKH	Java		9994
	GBO	Oxford, England		10000
	JBK	Kagoshima, Japan		10000
	——	Glace Bay, Canada		10000
	LCM	Stavanger, Norway (Proposed)		10000
	B2	Machelen les Brussels, Belgium		10000
	OCRU	Rufisque, Senegal, Fr. W. Africa		10000
	LGN	Norway		10000
	——	Mexico		10000
	——	France		10000
	2LA	Cuba		10000
10010	1XR	Manila, P. I.		10010
	PKH	Java		10005
	9AQ	Nipigon, Ont.		10020
	9AI	Toronto, Ont.		10020
10030	——	Spain		10030
	UOR	Vienna, Austria		10033
	——	Nauen, Germany		10040
10070	——	Nauen, Germany		10080
10110	DIT	Königs Wüsterhausen, Germany		10120
10150	DIS	Königs Wüsterhausen, Germany		10156
10170	KNR	Clearwater, Calif.	Mackay Radio & Teleg. Co.	10170

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
10870	GKQ	Dollis Hill, England		10870
	SQCN	Rio de Janeiro, Brazil		10870
	G5BH	England		10870
10890	KKC	Palo Alto, Calif.	Federal Telegraph Co.	10900
	——	Nauen, Germany		10900
	JYB	Tokyo, Japan		10900
	HI	Calgary, Alta.		10900
10910	KZED	Manila, P. I.		10903
	G5DH	Dollis Hill, England		10903
	VQG	Australia		10910
	ORU	Belgium		10910
10930	KGS	Honolulu, Hawaii	The Robt. Dollar Co.	10930
	KGR	Seattle, Wash.	The Robt. Dollar Co.	10930
	WGA	New York, N. Y.	The Robt. Dollar Co.	10930
	KGX	Los Angeles, Calif.	The Robt. Dollar Co.	10930
	KGQ	San Francisco, Calif.	The Robt. Dollar Co.	10930
	GLQ	Ongar, England		10930
10950	AIN	Morocco		10950
10970	CA	Geizers Hill, N. S.		10980
10990	——	Kootwijk, Holland		10990
	SABI	Goteborg, Sweden (Proposed)		10990
	SPW	Rio de Janeiro, Brazil		10990
11050	KRK	Palo Alto, Calif.	Mackay Radio & Teleg. Co.	11046
	WHD	New York, N. Y.	New York Times Co.	11040
11110	HI	Calgary, Alberta		11100
	AND	Java		11104
	PCMM	Kootwijk, Holland		11104
	SQBD	Recife, Pernambuco, Brazil		11110
	SQBC	Curityba, Parana, Brazil		11110
	SQBB	Porto Alegre, Brazil		11110
	——	Nauen, Germany		11110
	ICJ	Italy		11110
	JYZ	Tokyo, Japan		11110
	RCRL	Russia		11110
11170	RZA	Monrovia, Liberia		11170
	——	Königs Wüsterhausen, Germany		11180
11200	WKA	E. Pittsburgh, Pa.	Westinghouse Elec. & Mfg. Co.	11190
	OJO	Austria		11190
	WSL	Sayville, N. Y.	Mackay Radio & Teleg. Co.	11200
	WFE	(Portable)	Cmdr. Richard E. Byrd	11200
	WFD	(Portable)	Cmdr. Richard E. Byrd	11200
	KFK	(Portable)	Cmdr. Richard E. Byrd	11200
	WFA	(Portable)	Cmdr. Richard E. Byrd	11200
	WFB	Airplane *Floyd Bennett*	Cmdr. Richard E. Byrd	11200
	WFF	Airplane *Fokker*	Cmdr. Richard E. Byrd	11200
	WFC	Airplane *Fairchild*	Cmdr. Richard E. Byrd	11200
	HBC	Berne, Switzerland		11200
11260	——	Germany		11260
11290	GFA	London, England		11280
	UJ	Santa Marta, Colombia		11280
	AGB	München, Schleissheim, Germany		11280
	WFE	(Portable)	Cmdr. Richard E. Byrd	11300
	WFD	(Portable)	Cmdr. Richard E. Byrd	11300
	KFK	(Portable)	Cmdr. Richard E. Byrd	11300
	WFA	(Portable)	Cmdr. Richard E. Byrd	11300
	WFB	Airplane *Floyd Bennett*	Cmdr. Richard E. Byrd	11300
	WFF	Airplane *Fokker*	Cmdr. Richard E. Byrd	11300
	WFC	Airplane *Fairchild*	Cmdr. Richard E. Byrd	11300
11350	DAN	Norddeich, Germany		11340
11350	XOM	China		11365
11410	WKA	E. Pittsburgh, Pa.	Westinghouse Elec. & Mfg. Co.	11400
	WDJ	Harrison, Ohio	Crosley Radio Corp.	11400
	AFI	Königs Wüsterhausen, Germany		11410
	CG	Drummondville, P. Q.		11413
	GLY	Dorchester, England		11420
	——	Dorchester, England		11425
11440	ANC	Java		11440
	CF	Drummondville, P. Q.		11452
	——	Königs Wüsterhausen, Germany		11460
11500	——	New Zealand		11490
	GBK	Bodmin, England		11495
	——	Königs Wüsterhausen, Germany		11500

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
11530	CF	Drummondville, P. Q.		11531
	UF	Darrios, Guatemala	Tropical Radio Teleg. Co.	11540
	UB	Almirante, Panama		11540
	VIS	Sydney, Australia		11540
11560	VMG	Samoa		11560
	GBH	Grimsby, England		11575
11590	GBH	Grimsby, England		11580
	DIR	Königs Wüsterhausen, Germany		11580
11620	——	Königs Wüsterhausen, Germany		11620
	DFK	Nauen, Germany		11620
	FUT	Toulon, France		11621
	UB	Almirante, Panama		11630
11650	VIZ	Melbourne, Australia		11655
11680	VIZ	Ballan, Australian Commonwealth		11670
	KIO	Kahuku, Hawaii	Radio Corp. of America	11680
11710	——	Argentine		11720
	CJRX	Winnipeg, Manitoba		11720
11770	JYZ	Tokyo, Japan		11758
	GLKY	SS *Carinthia*, England		11758
	AGC	Nauen, Germany		11758
	AGC	München, Schleissheim, Germany		11760
	UL	Managua, Nicaragua		11780
11800	——	Königs Wüsterhausen, Germany		11800
	——	Vienna, Austria		11801
	OCTN	Toulon, France		11804
11830	——	Nauen, Germany		11840
11890	UL	Managua, Nicaragua		11880
	——	Königs Wüsterhausen		11900
11920	DGS	Nauen, Germany		11920
	ANC	Java		11920
11950	——	Spain		11940
	KKQ	Bolinas, Calif.	Radio Corp. of America	11950
11980	GBO	Leafield, England		11980
	LP	Buenos Aires, Argentine		11980
	FS1	Sainte Assise, France		11985
	CF	Drummondville, P. Q.		11990
	CG	Drummondville, P. Q.		11990
	CJ	Drummondville, P. Q.		11990
	CF	Drummondville, P. Q.		11992
	HZA	Saigon, French Indo-China		11993
	OCLY	Bordeaux, France		11993
12010	UR	San Jose, Costa Rica		12000
	JYZ	Tokyo, Japan		12000
	OCLY	Croix d'Hins, France		12000
	B82	Belgium		12000
	——	Quebec, Can.		12000
		Denmark		12000
	RZA	Monrovia, Liberia		12000
	VIY	Australia		12020
	VIY	Ballan, Australian Commonwealth		12025
12040	——		U. S. Government	12045
	FUT	France		12045
	——		U. S. Government	12051
12070	——		U. S. Government	12060
	——	Cuba		12070
	——		U. S. Government	12075
	FW	France		12085
12100	——		U. S. Government	12090
	ANC	Java		12090
	CJ	Drummondville, P. Q.		12093
	——	Recife, Pernambuco, Brazil (Proposed)		12094
	UR	Cartago, Costa Rica	Tropical Radio Teleg. Co.	12100
	——	Horomushiro, Japan (Proposed)		12100
	——	Quebec, Can.		12102
12130	——		U. S. Government	12135
	——	Manila, P. I.		12145
12160	——		U. S. Government	12150
	GBS	Rugby, England		12150
	FWX	Sainte Assise, France		12160
	——		U. S. Government	12165
12190	——		U. S. Government	12180
	FW	Sainte Assise, France		12188
	——		U. S. Government	12195
	JPM	Palaos, Japan		12200
12220	——		U. S. Government	12210
	——		U. S. Government	12225

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
12250	GLQ	Ongar, England		12240
	——		U. S. Government	12240
	——	Taihoku, Japan (Proposed)		12250
	——		U. S. Government	12255
12250	——	Dorchester, England		12265
12280	——	Spain		12270
	GBU	Rugby, England		12290
12310	JES	Osaka, Japan		12300
	JAN	Tokyo, Japan		12300
	——		U. S. Government	12315
12340	KFD	Denver, Colo.	General Elec. Co.	12338
	——	England		12350
12400	UJ	Santa Marta, Columbia	Tropical Radio Teleg. Co.	12400
	JRW	Parao, Japan		12400
	PKK	Dutch East Indies		12400
	——		U. S. Government	12405
12460	——		U. S. Government	12465
12490	GBM	Oxford, England		12492
	——	Oxford, England		12500
	JES	Osaka, Japan		12500
	——	Goteborg, Sweden (Proposed)		12500
	KCE	Liepaja, Latvia		12500
	9BE	Toronto, Ont.		12500
	5SW	Chelmsford, Can.		12500
	IRI	Rome, Italy		12500
	JEW	Japan		12500
	G5SW	England		12500
12520	——	Ontario, Can.		12511
	——	Quebec, Can.		12511
12610	——		U. S. Government	12615
12700	KHAH	Airplane *Greater Rockford*	Bert Hassell	12700
	——	Choshi, Japan (Proposed)		12700
	OLQ	Holland		12704
	——		U. S. Government	12705
12760	KTA	Guam, Mariana Island	Mackay Radio & Teleg. Co.	12758
	——		U. S. Government	12765
	PCH	Scheveningen, Holland		12766
12790	——		U. S. Government	12795
	JBD	Japan		12800
12820	JBD	Keijo, Japan		12825
	VCJ	Wakeham Bay, Ont.		12825
	VCJ	Ottawa, Ont.		12825
	CGP	Term Point, N. W. T.		12825
12850	3XK	Washington, D. C.	Jenkins Labs.	12840
	——		U. S. Government	12840
	1XAC	Providence, R. I.	C. E. Mfg. Co.	12850
12880	LAD	Norway		12880
	——		U. S. Government	12885
12910	——		U. S. Government	12900
	JFAB	Giran, Japan		12900
	——	France		12900
12940	WAX	Hileah, Fla.	Tropical Radio Teleg. Co.	12940
	WNN	Mobile, Ala.	Tropical Radio Teleg. Co.	12940
	UOX	Vienna, Austria		12931
12970	WNU	New Orleans, La.	Tropical Radio Teleg. Co.	12970
	WBF	Boston, Mass.	Tropical Radio Teleg. Co.	12970
13000	KNW	Palo Alto, Calif.	Mackay Radio Teleg. Co.	13000
13030	KNN	Honolulu, Hawaii	Mackay Radio Teleg. Co.	13030
	HZG	Numea, France		13043
	RAU	Moscow, Russia		13045
	RLT	Russia		13045
	PKH	Java		13045
	XON	China		13045
13090	——		U. S. Government	13095
13120	——		U. S. Government	13110
	——		U. S. Government	13125
13150	——		U. S. Government	13140
	——		U. S. Government	13155
	PCH	Scheveningen, Holland		13160
13180	DGG	Nauen, Germany		13180
	WFE	(Portable)	Cmdr. Richard E. Byrd	13187
	WFD	(Portable)	Cmdr. Richard E. Byrd	13187
	WFA	(Portable)	Cmdr. Richard E. Byrd	13187
	KFK	(Portable)	Cmdr. Richard E. Byrd	13187
	WFB	Airplane *Floyd Bennett*	Cmdr. Richard E. Byrd	13187
	WFF	Airplane *Fokker*	Cmdr. Richard E. Byrd	13187
	WFC	Airplane *Fairchild*	Cmdr. Richard E. Byrd	13187

Channel	Call Letters	Location of Transmitter	Owner	Frequency
14290	UG	Tegucigalpa, Spanish Honduras	Tropical Radio Teleg. Co.	14290
	VIA	Australia		14290
	VIB	Australia		14290
	VIT	Australia		14290
	VIS	Australia		14290
	VJZ	Australia		14290
	RKV	Russia		14290
	——	Dollis Hill, England		14290
14320	PKH	Soerabaja, Java		14311
14350	PKH	Java		14350
14410	——	Nauen, Germany		14400
14440	DIP	Königs Wüsterhausen, Germany		14437
	GBW	Rugby, England		14440
14470	WNC	Ocean Township, N. J.	American Tel. & Tel. Co.	14470
	——	Spain		14480
14500	PCLL	Kootwijk, Holland		14500
	——	Kootwijk, Holland		14500
	——	Scheveningen, Holland		14500
14560	UJ	Santa Marta, Colombia	Tropical Radio Teleg. Co.	14560
	——	Kootwijk, Holland		14560
	——	Scheveningen, Holland		14560
14590	WMI	Deal, N. J.	American Tel. & Tel. Co.	14590
	UQ	Bluefields, Nic.		14600
14620	——	Spain		14620
	PCRR	Sgravenhage, Holland		14625
	——	Japan		14630
	ANF	Java		14630
14680	——	Palo Alto, Calif.	Mackay Radio & Teleg. Co.	14680
	DFD	Nauen, Germany		14680
14710	UW	Cape Gracias, Nic.		14700
	9OB	Ottawa, Ont.		14700
	——	Palo Alto, Calif.	Mackay Radio & Teleg. Co.	14710
14740	——	Sayville, N. Y.	Mackay Radio & Teleg. Co.	14740
14770	GLSQ	SS *Olympic*, England		14769
	——	Sayville, N. Y.	Mackay Radio & Teleg. Co.	14770
14800	——	Nicaragua (Emergency)		14800
	WQV	Rocky Point, N. Y.	Radio Corp. of America	14800
14830	WKU	Rocky Point, N. Y.	Radio Corp. of America	14830
14860	WGA	New York, N. Y.	The Robt. Dollar Co.	14860
	KGS	Honolulu, Hawaii	The Robt. Dollar Co.	14860
	KGX	Los Angeles, Calif.	The Robt. Dollar Co.	14860
	KGR	Seattle, Wash.	The Robt. Dollar Co.	14860
	KGQ	San Francisco, Calif.	The Robt. Dollar Co.	14860
14890	1XS	Newton, Mass.	R. A. Fessenden	14890
	WGA	New York, N. Y.	The Robt. Dollar Co.	14890
	KGS	Honolulu, Hawaii	The Robt. Dollar Co.	14890
	KGX	Los Angeles, Calif.	The Robt. Dollar Co.	14890
	KGR	Seattle, Wash.	The Robt. Dollar Co.	14890
	KGQ	San Francisco, Calif.	The Robt. Dollar Co.	14890
	UL	Managua, Nic.		14900
14920	WAZ	New Brunswick, N. J.	Radia Corp. of America	14920
	PJN	Dutch Guinea		14920
14980	KGDQ	SS *Faith*, U. S. A.	Walden W. Shaw	14991
15010	1XS	Newton, Mass.	Howard A. Fessenden	15000
	C3BD	Santiago, Chile		15000
	C2BF	Vina del Mar, Chile		15000
	C3AC	Santiago, Chile		15000
	——	Kootwijk, Holland		15000
	——	Scheveningen, Holland		15000
	UF	Barrios, Guatemala		15000
	——	Taihoku, Japan (Proposed)		15000
	——	Norway (Proposed)		15000
	JBR	Siberia Maru, Japan		15000
	CF	Drummondville, P.Q.		15000
	CG	Drummondville, P. Q.		15000
	CJ	Drummondville, P. Q.		15000
	VIS	Australia		15000
	XGA	China		15000
	CH	Chile		15000
	GFR	England		15000
	JIPP	Japan		15000
	OCTN	Tunis, Tunisia		15000
	PJC	Curacao, Dutch W. Indies		15000
	RDRL	Russia		15000
	TUK	Russia		15000
15040	WQG	Rocky Point, N. Y.	Radio Corp. of America	15040
15070	DIO	Königs Wüsterhausen, Germany		15075
15100	2XBW	[1]Bound Brook, N. J.	Radio Corp. of America	15100

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[1] The frequencies between 15100 and 15200 kc have, in this case, been assigned as one band

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
15130	2XBW	[1]Bound Brook, N. J.	Radio Corp. of America	15130
15160	2XBW	[1]Bound Brook, N. J.	Radio Corp. of America	15160
15190	2XBW	[1]Bound Brook, N. J.	Radio Corp. of America	15190
	2XBW	[1]Bound Brook, N. J.	Radio Corp. of America	15200
15220	PCJJ	Eindhoven, Holland		15220
	—	Nauen, Germany		15220
15370	UR	Cartago, Costa Rica	Tropical Radio Teleg. Co.	15380
15430	KWE	Bolinas, Calif.	Radio Corp. of America	15430
15460	KKR	Bolinas, Calif.	Radio Corp. of America	15460
	WBF	Boston, Mass.	Tropical Radio Teleg. Co.	15460
15490	KEM	Bolinas, Calif.	Radio Corp. of America	15490
15520	VLW	New Zealand		15520
	AMF	Malabar, Java		15535
15550	UL	Managua, Nic.	Tropical Radio Teleg. Co.	15540
	UL	Managua, Nic.	Tropical Radio Teleg. Co.	15550
15580	—	Garden City, N. Y.	American Publishers Comm.	15580
	DFR	Nauen, Germany		15580
15610	—	Nauen, Germany		15600
	—	New York, N. Y.	American Publishers Comm.	15610
15640	—	Chicago, Ill.	American Publishers Comm.	15640
15670	—	New York, N. Y.	American Publishers Comm.	15670
15700	—	Floral Park, N. Y.	American Publishers Comm.	15700
	—	Argentine		15710
15730	—	San Francisco, Calif.	American Publishers Comm.	15730
15760	—	Boston, Mass.	American Publishers Comm.	15760
15790	FW	France		15780
	—	Taihoku, Japan (Proposed)		15800
	—	Nauen, Germany		15800
15820	DFN	Nauen, Germany		15820
15850	—	New York, N. Y.	American Publishers Comm.	15850
15880	—	Los Angeles, Calif.	American Publishers Comm.	15880
	BYB	Cleethorpes, England		15880
15910	—	Chicago, Ill.	American Publishers Comm.	15910
15940	DGI	Nauen, Germany		15940
	AND	Java		15950
15970	—	Nauen, Germany		15960
	WKO	Rocky Point, N. Y.	Radio Corp. of America	15970
16000	WKQ	Rocky Point, N. Y.	Radio Corp. of America	16000
16030	WKW	Rocky Point, N. Y.	Radio Corp. of America	16030
16060	—		U. S. Governemnt	16060
	—		U. S. Government	16068
16090	—		U. S. Government	16080
	—		U. S. Government	16100
	TQS	Lisbon, Portugal		16100
	KEB	Oakland, Calif.	General Elec. Co.	16102
16120	KEB	Oakland, Calif.	General Elec. Co.	16112
	—		U. S. Government	16120
16150	AYA	Venezuela		16140
16180	—		U. S. Government	16180
	—	Java		16180
16210	AGC	Nauen, Germany		16206
16270	WLO	Ocean Township, N. J.	American Tel. & Tel. Co.	16270
16300	—	Kootwijk, Holland		16300
	—	Kootwijk, Holland		16304
16330	—		U. S. Government	16320
	—		U. S. Government	16340
	—	Laonda, Angola		16340
	CRHA	Portuguese W. Africa		16340
16360	DIN	Königs Wüsterhausen, Germany		16358
16390	—	Manila, P. I.		16380
16420	—		U. S. Government	16420
	PQS	Lisbon, Portugal		16420
	PQS	Afragide, Portugal		16420
16460	WHD	New York, N. Y.	New York Times Co.	16460
	PJD	St. Martins		16460
	—	Goteborg, Sweden (Proposed)		16480
16500	CHRB	Cape Verde Island		16500
	CHRC	Angola		16500
16540	—		U. S. Government	16540
16580	CHRB	CapeVerde Island		16580
	CRHA	Portugese East Africa		16580
16620	—		U. S. Government	16620
16660	KTA	Guam, Mariana Island	Mackay Radio & Teleg. Co.	16660
	B2	Belgium		16660

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
16660	9CH	Geizers Hill, N. S.		16660
	DAN	Norddeich, Germany		16665
	——	Kootwijk, Holland		16670
	——	Nova Scotia		16670
	——	Riga, Latvia		16670
16780	WFB	Airplane *Floyd Bennett*	Cmdr. Richard E. Byrd	16800
	WFF	Airplane *Fokker*	Cmdr. Richard E. Byrd	16800
	WFC	Airplane *Fairchild*	Cmdr. Richard E. Byrd	16800
	WFA	(Portable)	Cmdr. Richard E. Byrd	16800
	WFD	(Portable)	Cmdr. Richard E. Byrd	16800
	WFE	(Portable)	Cmdr. Richard E. Byrd	16800
	KFX	(Portable)	Cmdr. Richard E. Byrd	16800
	WSL	Sayville, N. Y.	Mackay Radio & Teleg. Co.	16800
	RZA	Liberia		16800
16820	——		U. S. Government	16820
	SP	Brazil		16820
16860	WSL	Sayville, N. Y.	Mackay Radio & Teleg. Co.	16850
	GKT	Burnham, England		16845
	IDM	Italy		16850
16900	PJC	Curacao, Dutch West Indies		16900
	ORU	Belgium		16920
16940	KFD	Denver, Colo.	General Elec. Co.	16939
	——		U. S. Government	16940
16980	LR2	Italy		17000
17020	——		U. S. Government	17020
17060	——		U. S. Government	17060
17140	WKI	Newark, N. J.	Federal Telegraph Co.	17130
	——	Oxford, England		17140
	XGA	Mukden, China		17140
17180	——		U. S. Government	17180
	——		U. S. Government	17200
	PJD	St. Martins, D. W. I.		17200
17220	ANE	Java		17240
	GBO	Leafield, England		17240
17260	LPV	Monte Grande, Argentine		17241
	9XB	Culver, Ind.	L. E. Dutton	17270
17300	2XBY	Jersey City, N. J.	Walter C. Von Brandt	17300
	3XE	Baltimore, Md.	Baltimore Radio Show, Inc.	17300
	6XT	San Francisco, Calif.	C. L. Watson & R. L. Gray	17300
	1XAC	Providence, R. I.	C. E. Mfg. Co.	17300
	6XD	(Portable)	D. B. McGown	17300
17340	——	Nauen, Germany		17340
	DIM	Königs Wüsterhausen, Germany		17341
17380	LPI	Argentine		17380
17420	KNN	Honolulu, Hawaii	Mackay Radio & Teleg. Co.	17420
17460	——		U. S. Government	17460
	DGR	Nauen, Germany		17460
	——		U. S. Government	17480
	——	Königs Wüsterhausen, Germany		17480
17500	——		U. S. Government	17500
	AYF	Venezuela		17500
17540	——		U. S. Government	17540
	DFB	Nauen, Germany		17540
	VDB	Esquimalt, Canada		17560
17580	WBF	Boston, Mass.	Tropical Radio & Teleg. Co.	17580
17620	ANH	Malabar		17620
	KFVM	SS *Idalia*, USA	E. R. Parker	17636
	KKC	Palo Alto, Calif.	Federal Teleg. Co.	17640
17660	VJZ	New Guiana		17650
	G2BR	England		17650
	BQ	Machelen les Belgium		17650
	KNW	Palo Alto, Calif.	Mackay Radio & Teleg. Co.	17660
17700	WFE	(Portable)	Cmdr. Richard E. Byrd	17717
	WFD	(Portable)	Cmdr. Richard E. Byrd	17717
	WFA	(Portable)	Cmdr. Richard E. Byrd	17717
	KFK	(Portable)	Cmdr. Richard E. Byrd	17717
	WFB	Airplane *Floyd Bennett*	Cmdr. Richard E. Byrd	17717
	WFC	Airplane *Fairchild*	Cmdr. Richard E. Byrd	17717
	WFF	Airplane *Fokker*	Cmdr. Richard E. Byrd	17717
	——		U. S. Government	17720
17740	——		U. S. Government	17740
	——		U. S. Government	17744
17780	——	Kootwijk, Holland		17773
17820	PCRR	Kootwijk, Holland		17836
	——	Königs Wüsterhausen, Germany		17840
	——	Kootwijk, Holland		17840
17860	DIL	Königs Wüsterhausen, Germany		17850

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
19500	LP3	Buenos Aires, Argentine		19500
19540	DFQ	Nauen, Germany		19540
	—	Palo Alto, Calif.	Mackay Radio & Teleg. Co.	19540
19580	—	Sayville, N. Y.	Mackay Radio & Teleg. Co.	19580
	LP3	Monte Grande, Argentine		19600
19620	—	Sayville, N. Y.	Mackay Radio & Teleg. Co.	19620
	—		Radio Corp. of America	19620
	DFA	Nauen, Germany		19620
	GBO	Leafield, England		19628
19660	B82	Brussels, Belgium		19670
	—	Nauen, Germany		19680
19700	—		Radio Corp. of America	19700
	DFJ	Nauen, Germany		19700
19780	WTF	Akron, Ohio	Firestone Plantations Co.	19780
	WMU	Elizabeth, N. J.	Stan. Oil Co. of New Jersey	19780
	—		Radio Corp. of America	19780
	DGA	Nauen, Germany		19780
19820	WMI	Deal, N. J.	American Tel. & Tel. Co.	19820
19860	KZEN	Manila, P. I.	Radio Corp. of America	19860
	—	Nauen, Germany		19860
	2XV	Long Island City, N. Y.	Radio Engineering Labs.	19867
19900	LP4	Buenos Aires, Argentine		19900
19940	GLS	Ongar, England		19940
	DIH	Königs Wüsterhausen, Germany		19947
19980	CF	Drummondville, P. Q.		19987
	1XR	Manila, P. I.	Radio Corp. of America	19988
	GLS	Ongar, England		20000
	GKS	Dollis Hill, England		20000
	IR2	Rome, Italy		20000
	LCM	Stavanger, Norway (Proposed)		20000
	—	Glace Bay, Nova Scotia		20000
	—	Drummondville, P. Q.		20000
20020	AGA	Nauen, Germany		20020
	DFI	Nauen, Germany		20020
20060	AGA	Nauen, Germany		20055
	AGA	München Schleissheim, Germany		20070
	—	Nauen, Germany		20080
20100	—		U. S. Government	20085
	WQY	Rocky Point, N. Y.	Radio Corp. of America	20100
20140	—		U. S. Government	20125
	—	Nauen, Germany		20140
	—		U. S. Government	20150
20180	WQX	Rocky Point, N. Y.	Radio Corp. of America	20180
	—	Königs Wüsterhausen, Germany		20180
20220	—		U. S. Government	20225
	DIG	Königs Wüsterhausen, Germany		20230
20260	WQQ	Rocky Point, N. Y.	Radio Corp. of America	20260
20300	—	Sayville, N. Y.	Mackay Radio & Teleg. Co.	20300
20300	DIF	Königs Wüsterhausen, Germany		20312
20340	BYC	Horsea, England		20330
20380	—		U. S. Government	20400
	—	Nauen, Germany		20400
20420	DGB	Nauen, Germany		20420
20500	DGQ	Nauen, Germany		20500
	IR2	Italy		20500
20580	—	Rio de Janeiro, Brazil (Proposed)		20580
20620	6XU	Bolinas, Calif.	Radio Corp. of America	20620
20660	DGO	Nauen, Germany		20660
	LP2	Monte Grande, Argentine		20670
20700	FW	France		20700
20740	—	Königs Wüsterhausen, Germany		20730
	DGP	Nauen, Germany		20740
20780	KMM	Bolinas, Calif.	Radio Corp. of America	20780
20820	KSS	Bolinas, Calif.	Radio Corp. of America	20820
20860	—	Königs Wüsterhausen, Germany		20860
20900	BXC	Nore, England		20900
	LP3	Buenos Aires, Argentine		20900
20940	—	Nauen, Germany		20960
20980	FW	Sainte Assise, France		20966
	FW	Sainte Assise, France		20980
	—	Sayville, N. Y.	Mackay Radio & Teleg. Co.	20980
21060	WND	Ocean Township, N. J.	American Tel. & Tel. Co.	21060
21100	—	Königs Wüsterhausen, Germany		21100
21140	KZRC	Manila, P. I.	Radio Corp. of America	21140
21180	DGM	Nauen, Germany		21180
21220	WQA	Rocky Point, N. Y.	Radio Corp. of America	21220
21260	—	Königs Wüsterhausen, Germany		21260

Channel	*Call Letters*	*Location of Transmitter*	*Owner*	*Frequency*
21300	WQW	Rocky Point, N. Y.	Radio Corp. of America	21300
21340	DGM	Nauen, Germany		21340
21380	——	Sayville, N. Y.	Mackay Radio & Teleg. Co.	21380
	WBU	Rocky Point, N. Y.	Radio Corp. of America	21380
21420	WLO	Ocean Township, N. J.	American Tel. & Tel. Co.	21420
21740	——	Buenos Aires, Argentine		21740
21820	WFB	Airplane *Floyd Bennett*	Cmdr. Richard E. Byrd	21805
	WFF	Airplane *Fokker*	Cmdr. Richard E. Byrd	21805
	WFC	Airplane *Fairchild*	Cmdr. Richard E. Byrd	21805
	WFE	(Portable)	Cmdr. Richard E. Byrd	21805
	WFD	(Portable)	Cmdr. Richard E. Byrd	21805
	WFA	(Portable)	Cmdr. Richard E. Byrd	21805
	KFK	(Portable)	Cmdr. Richard E. Byrd	21805
22040	9CH	Geizers Hill, Nova Scotia		22050
22520	DGE	Nauen, Germany		22520
22580	DGF	Nauen, Germany		22600
22640	——		U. S. Government	22625
	KGQ	San Francisco, Calif.	The Robt. Dollar Co.	22660
	KGX	Los Angeles, Calif.	The Robt. Dollar Co.	22660
	WGA	New York, N. Y.	The Robt. Dollar Co.	22660
	KGS	Honolulu, Hawaii	The Robt. Dollar Co.	22660
	KGR	Seattle, Wash.	The Robt. Dollar Co.	22660
22800	WKA	East Pittsburgh, Pa.	Westinghouse Elec. & Mfg. Co.	22800
23780	LPZ	Monte Grande, Argentine		23809
25700	1XAC	Providence, R. I.	C. E. Mfg. Co.	25680
	3XK	Washington, D. C.	Jenkins Laboratories	25680
27260	GKS	Dollis Hill, England		27270
	G5DH	Dollis Hill, England		27270
27620	DID	Konigs Wusterhausen, Germany		27600
28580	XQM	Yunnanfu, China		28570
28880	DIC	Königs Wüsterhausen, Germany		28874
32240	2XBY	Jersey City, N. J.	Walter C. Von Brandt	34240
	3XE	Baltimore, Md.	The Balto. Radio Show Inc.	34240
	6XT	San Francisco, Calif.	C. L. Watson & R. C. Gray	34240
	1XAC	Providence, R. I.	C. E. Mfg. Co.	34240
	6XD	(Portable)	D. B. McGown	34240
51360	3XK	Washington, D. C.	Jenkins Laboratories	51360
	1XAC	Providence, R. I.	C. E. Mfg. Co.	51360

MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE*

THIS is a monthly list of references prepared by the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The articles listed below are not obtainable from the Government. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R000. RADIO COMMUNICATION

R060 Dellinger, J. H. The International Union of Scientific Radio Telegraphy. PROC. I.R.E., **16**, pp. 1107–12; Aug., 1928.

(Organization of U.R.S.I. and report of meeting of October 10–28, 1927.)

R100. RADIO PRINCIPLES

R110 Bramley, Arthur. Kerr effect in water due to high frequency radio waves. *Jnl.* Frank. Inst., **206**, pp. 151–158; Aug., 1928.

(Description of an experiment for finding the absorption lines of radio waves (3 to 300 cm) in water. A Kerr tube filled with water and placed between two nicol prisms and a quarter wave plate is used for the work.)

R113.4 Jouaust, R. Les Phenomenes de propagation des ondes radiotelegraphiques. (The propagation phenomena of radio waves.) *Comptes Rendus*, **187**, pp. 208–209; July 23, 1928.

(Calls attention to the fact that the propagation theory is based on a certain law for which the ionic density in the upper atmosphere is taken as a regular function of the altitude. The electrified particles given off from the sun are, however, a discontinuous emission similar to the Schottky effect in tubes.)

R113.7 Kenrick, G. W. Radio transmission formulae. *Phil. Mag.*, **6**, pp. 289–304; Aug., 1928.

(Derives a transmission formula for long-wave work taking the upper reflecting layer into account. The results indicate that the inverse square root of the wavelength in the damping factor of the Austin-Cohen formula has considerable theoretical justification, but the inverse square root of the distance should be used instead of the inverse first power of the distance. A slight change in the numerical constant is needed.)

R116 Frank, N. H. Die Fortplanzung elektrischer Wellen in Kabeln mit zwei Isolationsschichten. (The propagation of electric waves in cables with two insulating layers.) *Ann. d. Physik.*, **86**, pp. 422—34; June, 1928.

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* Original Manuscript Received by the Institute, September 14, 1928.

(The velocity of propagation of electric waves along cables with two insulating layers is computed according to Maxwell's theory. A new method is given for measuring the velocity along wires by means of using Lichtenberg-Figures. This gives a means for also determining the dielectric constant.

R116 Nancarrow, F. E. The behavior of a transmission line at radio frequencies. *Post Office Elec. Engrs. Jnl.*, **21**, pp. 165–69; July, 1928.

(Deals with the solution of the propagation of radio-frequency currents along a transmission line. Curves for the surge impedance for different spacings and size of wire are given.)

R125.1 Smith-Rose, R. L. Radio direction finder—the theory of the frame aerial avoiding electrostatic pickup. *Wireless World and Radio Review*, **23**, p. 186; August 15, 1928.

(Discussion of antenna effect and method of screening for its elimination using open loops.)

R127 Brunn, H. Eine einfache Methode zur Messung der Eigenwellenlänge von Antennen. (A simple method for the measurement of the natural wavelength of an antenna.) *Zeits. f. Hochfreq.*, **32**, p. 25; July, 1928.

(A tube wavemeter is loosely coupled to an antenna and the grid dip used for the determination of the natural wavelength.)

R130 Le Boiteux, H. L'influence des émissions secondaires des métaux sur le fonctionnement des lampes à trois électrodes. (Effect on the secondary emission of metals on the behavior of 3-electrode tubes.) Revue Gen. de L'Electricité, 28, pp. 939–46 and 984–992; June 2, 1928.

(This paper gives a theoretical investigation of the working characteristics of circuits using three-element tubes. The effect of secondary emission is taken into account.)

R130 Podlinsky, M. Equilibres instables et régimes statiques parasites dans les circuits électriques associés aux triodes. (Unstable equilibrium and disturbing conditions in triode circuits.) *L'Onde Electrique*, **7**, pp. 278–306; July, 1928.

(Theoretical treatment of the dynatron action in triode circuits.)

R130 Ballantine, Stuart. Schrot-Effect in high-frequency circuits. *Jnl.* Frank. Inst., **206**, pp. 159–168; Aug., 1928.

(A further theoretical study of the Schrot effect.)

R132.3 Hartshorn, L. Inter-electrode capacities and resistance amplification. *Experimental Wireless and Wireless Engineer*, 59, pp. 419–430; Aug., 1928.

(A study of resistance capacity coupled amplifiers with respect to distortion and interelectrode capacities and resistance.)

R133 Pfetscher, O. Ueber die Erregung sehr schneller elektrischen Schwingungen in der Dreielektrodenrohre. (On the production of oscillations of very high frequency by means of electron tubes.) *Phys. Zeits.*, **29**, pp. 449–478; July 15, 1928.

(Treats analytically the production of very short waves of tube oscillations and shows like Barkhausen and Kurz that the finite time for the electrons to pass to the respective electrodes produces a phase difference. The Gill-Morell oscillations are treated theoretically. The theory also explains the effect of the grid potential on the frequency even though the external circuit is kept unchanged.)

R133 Martyn, D. F. (A reply to K. E. Edgeworth.) Frequency variations of the triode oscillator. *Phil. Mag.*, **6**, pp. 223–228; July, 1928.

(Deals with the major effects causing a variation in the frequency of a triode oscillator. States that the most important cause of frequency variation is the flow of grid current, especially when the resistance is kept low and the frequency is not too high. Reference is made to generator with zero grid current which kept the frequency within one part in 100,000.)

R134 David, P. La détection par lampe. (Detection by means of electron tubes.) *L'Onde Electrique*, **7**, pp. 313–62; Aug., 1928.

(Theoretical and experimental data for the electron tube as a detector.)

R134.4 van der Pol, B. The effect of regeneration on the received signal strength. PROC. I. R. E., **16**, pp. 1045–52; Aug., 1928.

(Gives the theory and experimental verification of it for the effect of regeneration on signal strength.)

R140 Lion, K. Ein Wechselstromkompensator mit grossem Frequenzumfang. (An alternating current compensator with wide frequency range.) *Elekt. Nach.-Tech.*, **5**, pp. 276–83; July, 1928.

(A phase shifter for audio and high frequency currents is described.)

R142 Mallett, Prof. E. The resonance curves of coupled circuits. *Exp. Wireless and Wireless Engineer*, **59**, pp. 437–42; Aug., 1928.

(An analytical treatment (vectorial method) of coupled circuit with the frequency varied similar to the one described by the same author in the Feb. 1927 issue of *Experimental Wireless and Wireless Engineer* for single circuits.)

R154 Watanabe, Yasusi. Ueber die günstigste Belastung des Hochfrequenz-generators. (On the most favorable load for a high-frequency generator.) *Elekt. Nach.-Tech.*, **5**, pp. 259–267; July, 1928.

(A theoretical article for the best load to connect to a high-frequency alternator.)

R200. RADIO MEASUREMENTS AND STANDARDIZATION

R201 Clapp, J. K. A note on methods of rapidly adjusting a radio frequency oscillator in small steps of frequency. *Jnl.* Opt. Soc. of Amer., **17**, pp. 132–37; August, 1928.

(Describes rapid methods of adjusting the frequency of a generating set by aural methods using auxiliary generating sets and a tuning fork. Discusses accuracy obtainable.)

R201.6 Ferguson, J. G., and Bartlett, B. W. Measurement of capacitance in terms of resistance and frequency. *Bell System Tech. Jnl.*, **7**, pp. 420–37; July, 1928.

(Describes an adaptation of the Wien Bridge for the determination of capacity.)

R214 Wheeler, L. P., and Bower, W. E. A new type of standard frequency piezo-electric oscillator. PROC. I. R. E., **16**, pp. 1035–44; August, 1928.

(Gives a new type of piezo-electric oscillator, for which the oscillations are sustained by an acoustic feedback.)

R214 Hund, A. Notes on quartz plates, air gap effect, and audio-frequency generation. PROC. I. R. E., **16**, pp. 1072–78; August, 1928.

(The effect of supersonic sound waves taking place in a crystal holder is discussed. The energy curve is given with respect to the air gap in the crystal holder. van der Pol's relaxation oscillations are suggested in the circuit for obtaining very accurate low-frequency current from a high-frequency crystal.)

R220 Griffiths, W. H. F. The measurement of small variable capacities at radio frequencies. *Experimental Wireless and Wireless Engineer*, 59, 452–59; August, 1928.

(A new method is given for calibrating small capacities.)

R223 Bryan, A. B., and Sanders, I. C. The dielectric constant of air at radio frequencies. *Physical Review*, **32**, pp. 302–10; August, 1928.

(Value of dielectric constant of dry air free from carbon dioxide determined as 1.0005893 for standard conditions of temperature and pressure. The method is a modification of the usual heterodyne beat arrangement.)

R290 Obata, Juichi. The "Ultramicrometer," a new instrument for measuring very small displacement or motion, and its various applications. *Jnl.* Opt. Soc. of Amer., **16**, pp. 419–32; June, 1928.

(Describes the construction and various examples of applications of the ultramicrometer for measuring a small displacement or motion utilizing a generating electron-tube circuit. The displacement or motion to be measured is made to produce either a change in capacity or in the eddy current loss, and in consequence a corresponding change in the plate current of the tube.)

R300. RADIO APPARATUS AND EQUIPMENT

R342 Jouaust, R., and Decaut, B. Note sur quelques perfectionnements des amplificateurs a courant continu. (Notes on some perfections on d.c. amplifiers.) *L'Onde Electrique*, **7**, pp. 306–08; July, 1928.

(A straight two-stage d.c. amplifier is used with the first tube having a large amplification factor. In one case a special tube is employed in the first stage with a narrow mesh grid and a large anode ($\mu = 36$). In another case a double-grid tube is used with the two grids connected together.)

R342 Hund, A. Notes on aperiodic amplification and applications to the study of atmospherics. PROC. I. R. E., **16**, pp. 1077–78; August, 1928.

(A circuit for aperiodic amplification is given and the main differences between aperiodic amplification and amplification of harmonic voltages are brought out.)

R342.7 Reppisch, H. Ueber die Spannungsverstärkung mittels Transformatorenkopplung beim Niederfrequenzverstärker. (On voltage amplification by means of transformer coupling for the low-frequency amplifier.) *Zeits. f. Hochfrequenz.*, **32**, pp. 22–24; July, 1928.

(A derivation of the relations for transformer coupled amplification (audio frequency).)

R342.7 Kirke, H. L. Microphone amplifiers and transformers. *Experimental Wireless and Wireless Engineer*, 59, pp. 443–51; August, 1928.

(A continuation of the article appearing on p. 370 in the July issue. Deals with the effects of coupling between stages of an amplifier. Takes inter-electrode capacity into account.)

R342.7 Thomson, J. M. Characteristics of output transformers. PROC. I. R. E., **16**, pp. 1053–64; August, 1928.

(Discusses analytically and by means of tests, the characteristics of output transformers.)

R343.7 Kimmell, W. J. The cause and prevention of hum in receiving tubes employing alternating current direct on the filament. Proc. I. R. E., **16**, pp. 1089–1106; August, 1928.

(The effect of the filament which is heated with a.c. on the quality of reception is discussed.)

R344 Bell, Eric G. A valve-maintained high-frequency induction furnace and some notes on the performance of induction furnaces. *Proc.* Phys. Soc. (London), **40**, pp. 193–205; June 15, 1928.

(Gives electrical design of an induction furnace and gives a general theory of the behavior of induction furnaces. Experimental results given supporting the theory.)

R344 Hund, A. Générateur des courants de fréquence audible et réglable a stabilisation piézo-electrique. (Generator for audio currents of adjustable frequency with piezo-electric stabilization.) *QST Francais*, **9**, pp. 16–19; August, 1928.

(A translation of Bureau of Standards Scientific Paper No. 569, giving a method of producing audio currents of variable frequency and good wave form.)

R344.4 Wechsung, H. Röhrengenerator grosser Leistung für sehr kurze elektrische Wellen. (Tube generator of large rating for very short waves.) *Zeits. f. Hochfreq.*, **31**, pp. 176–83; June, 1928.

(A circuit is discussed by means of which 700 watts energy for code modulation and 300 watts energy for speech modulation can be generated. Waves down to 4.2 meters were produced.)

R344.4 Hollmann, H. E. Telephonie auf extrem kurzen Wellen. (Telephony with extremely short waves.) *Elekt. Nach.-Tech.*, **5**, pp. 268–75; July, 1928.

(The author describes a system for telephony with wavelengths between 30 and 100 cm. It is also shown that the Barkhausen and Gill and Morrell oscillations can occur simultaneously.)

R344.6 Lübcke, E. Eine Gross-Verstärkerröhre mit Quecksilberdampf. (A power amplifier tube with mercury vapor.) *Zeits. f. Hochfreq.*, **32**, pp. 1–10; July, 1928.

(An amplifier tube using a mercury arc is described. The internal resistance is only 70 ohms and the tube seems promising for heavy current work.)

R374.1 Ogawa, W. Analogy between the crystal detector and a vacuum tube. *Phil. Mag.*, **6**, pp. 175–78; July, 1928.

(The author explains the rectification of a crystal detector by means of a difference of electrons emitted from each electrode. According to his views there is no substantial difference between a crystal detector and a vacuum-tube rectifier except the metallic conduction at the real contact points in the former.)

R374.1 Regler, Fritz. Vorläufige Mitteilung uber die Theorie des Kontaktdetektors. (Preliminary communication on the theory of contact detectors.) *Phys. Zeits.*, **29**, pp. 429–36; July, 1928.

(The author divides contact rectifiers into two classes. For the first class rectification is due to electrostriction (piezo-electric effects); and for the other, due to different values of electron affinity for different materials. Many of his conclusions are based on a paper by G. G. Reisshaus, *Phys. Zeits.*, 28, 223; 1928.)

R374.1 Beck, P. Weitere Mitteilungen zum Kristalldetektor problem. (Further communication on contact detectors.) *Phys. Zeits.*, **29**, pp. 436–37; July, 1928.

(Describes experimental work with galena-silver, galena-copper and galena-stee contact rectifiers. Microscopic observations of the contact surface have been made in addition, and the results seem to be in agreement in several ways with the ones due to G. G. Reisshaus.)

R381 Trogner, A. M. Mica condensers for high frequency. *QST*, **12**, pp. 47–49; September, 1928.

(Gives method of connecting small mica condensers for use in high-frequency transmitting sets.)

R388 Lee, E. S. Cathode-ray oscillographs and their uses. *General Electric Review*, **31**, pp. 404–12; August, 1928.

(Describes principles of cathode-ray oscillograph and commercial instrument, in which photographic records can be made. Illustrations are given for its use on various problems including radio.)

R400. RADIO COMMUNICATION SYSTEMS

R432 Jones, L. J., and Osborn, W. M. Humber radio station. *Post Office Elec. Engrs. Jnl.*, **21**, pp. 159–64; July, 1928.

(Description of station located at Mablethorpe, Lincolnshire.)

R470 Dubois, R. Installation télégraphique à haute fréquence avec appareils "Télétype" réalisée sur une lique de transmission d'énergie a 60,000 volt de la Société d'Electricité du Tarn. (Telecommunication over the 60,000-volt transmission line of the Société d'Electricité of Tarn.) *Revue Gen. de L'Electricite*, **28**, pp. 997–1003; June 9, 1928.

(Describes the installation of the "Télétype" system for sending messages over a high-voltage line. The line is 90 km long and rated at 60 kilovolts. The system works entirely automatically and gives the messages directly in ordinary writing.)

R500. APPLICATIONS OF RADIO

R522 Krueger, K., and Plendl, H. Zur Anwendung der kurzen Wellen im Verkehr mit Flugzeugen: Versuche zwichen Berlin und Madrid. (On the application of short waves to aeroplanes: Experiments between Berlin and Madrid.) *Zeits. f. Hochfreq.*, **31**, pp. 169–76; June, 1928.

(It was found that 46 m would work over a distance of 1400 km during daytime with 2 watts output, but a distance of 2000 km could hardly be covered even if 300 watts were used. The wavelength between 27 and 30 m seemed to work best during day and night time for a distance of 2000 km. 300 watts energy in the antenna would give a sure communication on ground. For flights, 30 watts gave a fairly dependable service. The band between 16 and 19 m was nearly as good. It was generally found that 300 watts and 30 m wavelength was dependable.)

R536 Experiments in underground communication through earth strata. U. S. Bureau of Mines Technologic Paper No. 433.

(Includes data on radio methods applied to such communication.)

R582 Larner, E. T. Practical television (book). Published by Ernest Benn, Ltd., London.

(Reviewed in *Nature*, **122**, No. 3068, Aug. 18, 1928.)

R800. NON-RADIO SUBJECTS

517 Berg, Ernst J. Heaviside's operational calculus as applied to engineering and physics. *General Electric Review*, **31**, pp. 444–51; August, 1928.

(Reviews Heaviside's operational calculus for the case of the asymptotic solution.)

534 Hubbard, J. C., and Loomis, A. L. The velocity of sound in liquids at high frequencies by the sonic interferometer. *Phil Mag.*, **5**, pp. 1177–90; June, 1928.

(A method is described where the velocity of sound in liquids is determined by means of standing waves generated by a quartz oscillator producing high-frequency vibrations. Thermodynamic coefficients of liquids are computed from the velocities obtained at different temperatures.)

534 Hehlgans, F. W. Uber Piezoquarzplatten als Sender und Empfänger hochfrequenter akustischer Schwingungen. (On piezoelectric quartz plates as sender and receiver of supersonic sound waves.) *Annalen der Physik*, No. 12, **86**, pp. 587–627; 1928.

(Experimental investigation of vibrating quartz plates for acoustic work.)

534 Trendelenburg, F. Zusammenfassender Bericht. Uber neurere akustische und insbesondere elektroakustische Arbeiten. (Summary of electro-acoustic methods.) *Zeits. f. Hochfreq.*, **32**, pp. 27–34; July, 1928.

(A compilation of acoustic methods used in a radio laboratory.)

535.3 Wynn-Williams, C. E. The application of a valve amplifier to the measurement of x-ray and photo-electric effects. *Phil. Mag.* **6**, pp. 324–34; August, 1928.

(The author applies a two-electron tube bridge circuit to the measurement of x-ray and photoelectric effects.)

546.432 Schindelhauer, F. Radioaktive Niederschläge auf Hochantennen (Radioactive effects on antennas.) *Phys. Zeits.*, **29**, pp. 479–87; July 15, 1928.

(Uses antennas to study atmospheric potentials. It is shown that the vertical current is mostly due to a radioactive deposit. The current increases with increasing air pressure and with increasing temperature of the ground.)

621.313 Liwschitz, M. Verhalten des selbsterregten Generators bei kapazitiver Belastung. (Control of the speed of d.c. motors by means of electron tubes.) *Wissen. Veroff. aus dem Siemens Konzern*, **6**, pp. 23–25; 1927.

(The voltage which is proportional to the r.p.m. is applied to a regular electron tube. The plate current of the latter affects the field of the motor.)

621.385 Küpfmüller, Karl and Mayer, Hans F. Sur les phénomenès transitoires dans les lignes pupinisées et le moyen d'y remédier. (Propagation of signals along conductors using Pupin coils.)

(Deals in detail with phase distortion and suggests filters for overcoming such distortion. Theoretical and experimental data prove their usefulness.)

621.385 Affel, H. A., Demarest, C. S., Green, C. W. Carrier systems of long distance telephone lines. *Bell System Technical Jnl.*, **7**, pp. 564–629; July, 1928.

(This paper is a continuation of the development of carrier systems on long distance telephone lines.)

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	Charlotte, c/o Radio Sta. WBT	Reid, Ralph J.
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	Tyko, Electro Technical Lab	Matsumura, Sadao
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New York	Buffalo, 39 Charleston Av	Patterson, Curtis B.
Pennsylvania	Philadelphia, 1533 Pine St	Gilson, Walter E.
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APPLICATIONS FOR MEMBERSHIP

Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Committee on Admissions. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before December 1, 1928. These applicants will be considered by the Board of Direction at its December 5th meeting.

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Japan	Tokyo, Setagaya, 429 Taishido	Fujimoto, Tadashi

For Election to the Associate grade

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California	Glendale, 408 Wing Street	Shomler, H. B.
	Hollywood, 1030 N. Western Avenue	Finstein, Paul
	Hollywood, 1345 North New Hampshire Avenue	Sarver, Frank M.
	Los Angeles, 4503 Paulhan	Andersen, H. W.
	Los Angeles, 1444 So. Norton Avenue	Buffinger, William
	Los Angeles, 1441 W. 51 Place	Hill, Alfred LeRoy
	Los Angeles, 1050½ Cloverdale	Stevens, Sterling M.
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	Stockton, 1520 N. Eldorado Street	Holt, Pliny E.
Georgia	Atlanta, c/o General Electric Co., P.O. Box 1698	Bussey, H. E.
	Atlanta, 1331 Lucile Avenue, S. W.	Wallace, Lawrence T.
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	Chicago, 2247 Calumet Avenue	Hauser, Albert
	Chicago, c/o R. C. A., 100 W. Monroe Street	Kennedy, W. J. B.
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	Chicago, 4430 Clifton Avenue	Turner, C. G.
	Dixon, 521 McKenney Street	Hall, Howard J.
	Kewanee, 219½ West Second Street	Larson, Clifford L.
	Urbana, Dept. of Elec. Eng., University of Illinois	Hershey, Arthur W.
Indiana	Indianapolis, 1036 Eugene Street	Byers, Russell R.
	Valparaiso, 402 Monroe Street	Alexander, Leslie Alpheus
	Valparaiso, 555 So. College Avenue	Imler, Glenn M.
Iowa	Hale, Box 29	Kruse, Gerald D.
Louisiana	Baton Rouge, 830 North Street	Webre, Andrew S.
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Massachusetts	Cambridge, Massachusetts Institute of Technology	Barrow, W. L.
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	Toronto, Ont., 139 Sunny Side Avenue	Cormack, William A.
	Toronto 2, Ont., Canadian Westinghouse Co., Ltd.	Thompson, Frank
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England	Blackford, Lancs, 6 Springfield Road	Cohen, Samuel H.
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PROCEEDINGS OF

The Institute of Radio Engineers

Volume 16 December, 1928 Number 12

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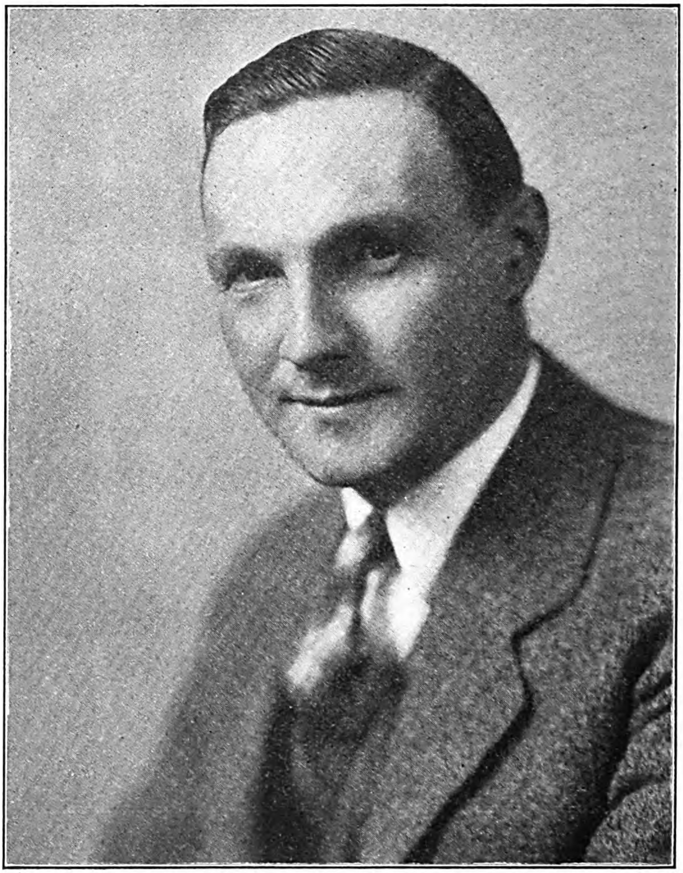

Alfred H. Grebe
Manager of the Institute, 1928

Alfred H. Grebe

MEMBER OF THE BOARD OF DIRECTION OF THE INSTITUTE, 1928

Alfred H. Grebe was born on April 4, 1893 at Richmond Hill, Long Island, N. Y. He graduated from high school and attended the American Wireless Institute. In 1907 Mr. Grebe began his experimental work in radio communication. From 1909 to 1912 he was employed as an operator by the United Wireless Telegraph Company. In 1912 he became associated with the Telefunken Company and the Atlantic Communication Company, with whom he served until 1914. Early in 1915 Mr. Grebe began manufacturing experimental radio apparatus. In 1916 he joined the staff of Kilbourn and Clark Company as a construction engineer.

In 1917 Mr. Grebe began manufacture and development of regenerative receivers. He was one of the earliest manufacturers of radio receivers in the country.

In connection with his work with the Telefunken Company, he was in charge of installation of twenty-seven stations. With Kilbourn and Clark Company he had active charge of construction and installation of fourteen stations. He also was associated with the rearrangement of the old Sayville Transatlantic Transmitter, located on Long Island.

In 1919 he became President of the A. H. Grebe Company, Inc., and has served in that capacity to date. Mr. Grebe was elected an Associate member of the Institute in 1917, and was transferred to the Fellow grade in November, 1927.

Mr. Grebe served as a member of the Board of Direction of the Institute from 1924 through 1926, and was reappointed a member of the Board in January, 1928.

Throughout his long active association with the Institute and its various activities, Mr. Grebe has served on practically all of its committees.

CONTRIBUTORS TO THIS ISSUE

Bailey, Austin: Received A.B. degree, University of Kansas, 1915; Ph.D., Cornell University, 1920; assistant and instructor in physics, Cornell University, 1915–18; signal corps, U. S. A., 1918–19; Fellow in physics, Cornell University, 1919–20; Corning Glass Works, 1920–21; assistant professor of physics, University of Kansas, 1921–22; Department of Development and Research, American Telephone and Telegraph Company, 1922—. Dr. Bailey's work while with the American Telephone and Telegraph Company has been largely along the line of methods for making radio transmission measurements and long-wave radio receiving problems. Associate member, Institute of Radio Engineers, 1925; Member, 1925.

Ballantine, Stuart: (See PROCEEDINGS for February, 1928).

Dean, S. W.: Received A.B. degree, Harvard University, 1919; Cutting and Washington, Inc., 1917; ensign, U. S. Naval Reserve Force, 1918; Radio Corporation of America, 1919–25; Department of Development and Research, American Telephone and Telegraph Company 1925—. Mr. Dean's work has been chiefly in connection with long-wave transatlantic radio receiving systems. Junior member, Institute of Radio Engineers, 1914; Associate, 1918; Member, 1926.

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Koehler, Glenn: Born November 20, 1894, at Van Wert, Ohio. Received B.S. degree from University of Illinois, 1918. M.S., University of Wisconsin. Radio research, signal corps, U. S. Army at Camp Alfred Vail, N. J., 1918. Research Department, Western Union Telegraph Company, 1919. Department of Electrical Engineering, University of Wisconsin, 1920 to date. Associate member, Institute of Radio Engineers, 1927.

Landon, V. D.: Born May 2, 1901. Educated Detroit Central High School and Central College. Engineering Department, Westinghouse Electric and Manufacturing Company, 1922—. At present in charge of Receiver Test Laboratory and Radio Frequency Development Laboratory. Associate member, Institute of Radio Engineers, 1927.

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Wolff, Irving: Received B.S. degree, Dartmouth College, 1916; nontechnical work, 1916–19; instructor, Iowa State College, 1919–20; instructor, Cornell University, 1920–23; Ph.D., Cornell University, 1923; research on polarization capacity for Heckscher Research Council, 1923–24; research on electro-acoustics, Radio Corporation of America, 1924. Author of a number of papers published in various technical magazines. Member, American Association for Advancement of Science and American Physical Society; member honorary societies, Sigma Xi and Phi Kappa Phi; Associate member, Institute of Radio Engineers, 1927.

The following are biographies of authors whose papers appeared in the October, 1928 issue of the PROCEEDINGS:

Montefinale, Gino: Born June 9, 1881. Educated at Technical Colleges in Italy; entered Naval Academy, London, 1899; officer of the Royal Navy, 1903; radio instructor at the R. N. Telegraphist School, 1908–10; assisted in erection of the high-power Marconi station at Mogadiscio, 1911; radio officer, Red Sea flotilla during Turkish-Italian war; director of radio service in Italian Somaliland, 1912–14; served in dreadnaught squadron during World War; sent to Red Sea as director of the Erytrea radio service, December, 1916; remained at Marconi station of Massawa till armistice; chief of radio laboratory in the dockyard of La Spezia, June, 1919 to April, 1925; chief of radio section of the Italian Navy Department and commander of the S. Paolo (IDO) station, October, 1925 to date. Italian delegate at the International Radiotelegraph Conference of Washington, 1927. Author of several papers on radio subjects and active correspondent of various periodicals and reviews. Member of Specialist Body of the Italian Navy, and member of the Superior Committee of Vigilance on Italian Broadcasting Services. Commander, Royal Italian Navy.

Pession, Giuseppe: Born May 30, 1881. Educated at Royal Naval Academy; appointed midshipman in 1902; director of La Spezia Radiotelegraphic School and teacher of radiotelegraphy at Rome Military Institute of Radiotelegraphy; since 1920 professor of radiotelegraphy and naval magnetism at Naples Royal Superior Polytechnical School; member of several commissions dealing with development and reorganization of the radiotelegraphic services both in Italy and abroad; erection of Rome-S. Paolo station and other radio installations; chief of radio services in Italian Navy Department and commander of Rome radio stations, 1917–24; awarded professor's degree in electrical and radio sciences, Polytechnical School of Naples and Rome, 1924; chief of Italian Postal and Telegraph Administration, 1925 to date. Author of scientific books and active correspondent of scientific magazines. Member of Specialist Body of the Italian Navy; vice-president of radio section, Consiglio Nazionale Delle Ricerche. Captain, Royal Italian Navy.

INSTITUTE NOTES AND RELATED ACTIVITIES

November Meeting of the Board of Direction

The regular meeting of the Board of Direction of the Institute was held in the offices of the Institute in the Engineering Societies Building, New York City, at 2 P.M., November 7th. The following were present: Alfred N. Goldsmith, President; L. E. Whittemore, Vice-President; Melville Eastham, Treasurer; Ralph Bown, Junior Past President; Arthur Bacheller, W. G. Cady, J. H. Dellinger, R. A. Heising, R. H. Manson, R. H. Marriott, C. J. Porter, Assistant Secretary, and John M. Clayton, Secretary.

Acting upon the recommendation of the Committee on Admissions, the Board approved the transfer or election of the following to higher grades of membership:

Transferred to the grade of Fellow: Carl Dreher. Transferred to the grade of Member: Frederick H. Drake, Harold Hardy, Charles T. Manning, and F. W. Van Why. Elected to the grade of Member: Thomas McL. Davis, J. Barton Hoag, and Charles C. Kolster.

Seventy-six Associate members and fourteen Junior members were elected.

Liebmann Memorial Prize Awarded to Dr. Cady

At the New York meeting of the Institute, held in the Engineering Societies Building on the evening of November 7, the 1928 award of the Morris Liebmann Memorial Prize was made to Walter G. Cady, of Scott Laboratory, Wesleyan University.

In presenting the prize, President Alfred N. Goldsmith, of the Institute, made the following remarks:

> (President Goldsmith)—In pursuance of the normal duties of the Institute, at this time it becomes my pleasant duty to announce the award of the Morris Liebmann Memorial Prize. This prize is awarded this year to Dr. Walter G. Cady, Head of the Department of Physics of Wesleyan University, for his fundamental investigations in piezo-electric phenomena and their applications to radio technique. With the nature of the work of Dr. Cady, of its importance, its influence on the development of the art, you

are all acquainted. With the progress that has resulted from these applications, you are also acquainted.

I have here the duly engrossed check which is at once a sentimental and material indication of the award in question. Dr. Cady, it is my great pleasure to present to you this check.

(Walter G. Cady)—Mr. Chairman, I appreciate very deeply this honor that you and the Directors of the Institute have conferred upon me. At various times and places I have been present when prizes were awarded, but I find now that the memory of having been a witness of the embarrassment of others is but very little help in knowing just what to say on this joyous occasion.

It was some years ago that I had occasion to measure the capacities and dielectric constants of certain Rochelle salt plates. I mention this because it is a good illustration of a principle that I have for a good many years felt to be rather an important one in experimental work, namely, that whenever you run across anything that seems to be an obstacle, or a source of error in your work, you can't do anything better than try to turn that to your advantage—just as all through life perhaps we try to make benefits so far as possible out of our various trials and tribulations. At any rate, the principle worked in this particular case, for what was at the time a very annoying source of error, as I thought, in carrying out those measurements, turned out to be the beginning of the series of experiments that have led to the applications of piezo-electricity in radio as we have them today.

I shall not attempt to trace out the development of that at all, but simply mention it as an instance of the advisability of being on the watch for opportunities to turn to your advantage those things that of themselves tend to be only obstacles to success.

At that time, when the first papers on this subject were published, I think, Mr. Chairman, that the Board of Directors would have been less inclined to bestow a medal than free transportation to the nearest insane asylum because the practical value of piezo-electricity seemed to be dubious and very remote. Since then the development that has taken place—very largely contributed to, I ought to say, by Professor Pierce, Dr. Miller, Commander Taylor, and others—has brought the use of piezo-electric crystals to the position that they hold today.

Again, Mr. Chairman, I want to thank you for your kindness, and for this honor.

Forthcoming Papers

The following is a partial list of papers on hand for publication in the Proceedings. These papers will probably appear in early forthcoming issues:

Fading Curves and Weather Conditions—by R. C. Colwell.

Reception Experiments in Mt. Royal Tunnel at Montreal, Quebec—by A. S. Eve, W. A. Steel, G. W. Olive, A. R. McEwan, and J. H. Thompson.

On the Behavior of Networks with Normalized Meshes—by E. A. Guillemin and W. Glendinning.

A Direct Reading Radio-Frequency Meter—by R. C. Hitchcock.

On the Mechanism of Electron Oscillations in Three-Electrode Tubes—by H. E. Hollman.

An Auxiliary Frequency Control for R. F. Oscillators—by G. F. Lampkin.

Further Researches on Determination of the Most Favorable Radiation Angle with Horizontal Polarization—by Alexander Meissner.

The Radiation Resistance of Beam Antennas—by A. A. Pistolkors.

A Method of Treating Resistance Stabilized Radio-Frequency Amplifying Circuits—by B. L. Snavely and J. S. Webb.

Detection Characteristics of Three-Element Vacuum Tubes—by Frederick E. Terman.

The Piezo-Electric Crystal Oscillator—by J. Warren Wright.

A Note on the Directional Observations on Grinders in Japan—by E. Yokayama and T. Nakai.

Byrd Antarctic Expedition

In connection with the account of some of the radio activities of the Byrd Antarctic Expedition, appearing in the Institute Notes Section of the PROCEEDINGS for October, an unfortunate omission of the name of Howard F. Mason as one of the members of the radio personnel occurred. Mr. Mason is well known for his previous radio experience with Captain Wilkins in the Arctic Regions, was responsible for the design and construction of much of the radio apparatus of the expedition, and supplied a considerable portion of the facts from which the above mentioned material was formulated for the PROCEEDINGS. He accompanies the expedition in the capacity of radio operator and observer.

1929 Convention

The annual Convention of the Institute will be held in 1929 in Washington, D. C. on May 13th to 15th inclusive, instead of early in January as has been the custom in the past several years.

The tentative Convention plans call for technical sessions and trips during the first two days with a banquet on the evening of the second day. The third day of the Convention will be devoted to participation in the annual meeting of the U.R.S.I.

Complete announcement of the detailed arrangements for the Convention will be published in an early forthcoming issue of the PROCEEDINGS.

It is suggested that members of the Institute arrange their plans to reserve the three days in May of 1929 for the Convention, as it is anticipated that an extremely interesting program will be presented.

Applications for Membership

The signatures of references for application for transfer or election to any grade of membership in the Institute are not required. Members are requested to so advise prospective applicants in order that such applications may be transmitted promptly to the Institute for consideration by the Committee on Admissions and the Board of Direction.

Changes of Address

The PROCEEDINGS of the Institute are sent to the membership under fourth class mail. The postal regulations do not allow fourth class mail to be forwarded when the addressee has removed from the address given. Such mail, when undelivered to the addressee, is returned by the post office to the Institute office. For this reason members of the Institute are *urged* to notify the Institute office promptly of any change in their mailing address so that copies of the PROCEEDINGS may reach them promptly. A change of address to be effective with a particular issue must reach the Institute office by not later than the fifteenth of the month preceding the date of publication. That is, a change of address to be effective with the October issue must reach the Institute office not later than September 15th.

In notifying the Institute office of change of address, members are requested to indicate not only their change in mailing address but their change in business connection, if any.

The *Graf Zeppelin* Radio Equipment

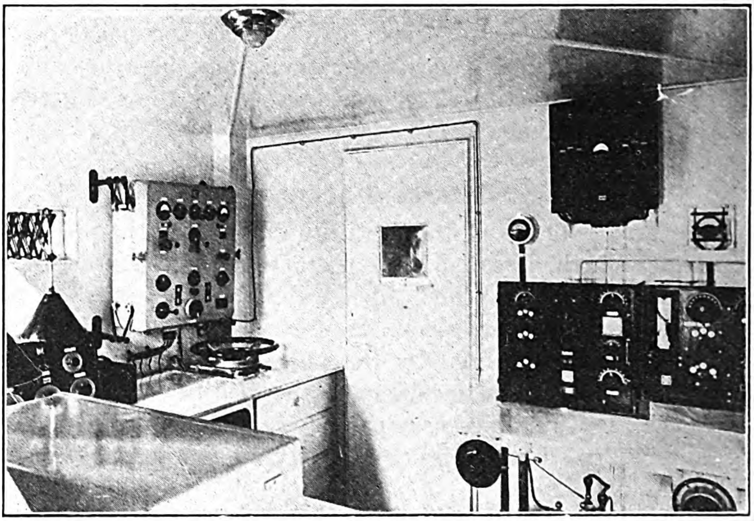
Through the courtesy of the Navy Department, we have obtained the following facts regarding the radio equipment and wavelengths used by the *Graf Zeppelin* in its recent transatlantic flight. This data was supplied by the Zeppelin Company of Germany.

The following wavelengths were assigned to the *Graf Zeppelin* whose call letters were DENNE.

Meters	
1875 1887 1911 1935 1961	Wavelengths only for communication with shipboard stations which can relay telegrams to coast stations.
1987	Sending wave for traffic with shipboard stations.
2098	General call wave of the airship for shipboard and coast stations.
2479 2521	Emergency call wave for European coast stations.

In addition waves 1400 and 900 meters for traffic with other aircraft.

The main transmitter was a 140-watt combination telephone and telegraph outfit having normal ranges of 1500 kilometers for telegraph and 400 kilometers for telephone.



Radio Room on the *Graf Zeppelin*

An emergency vacuum-tube transmitter with a rated output of 70 watts was also supplied. The normal operating range of the emergency set was 750 kilometers for telegraph and 180 kilometers for telephone.

Both transmitters were arranged for straight continuous-wave telegraph as well as for modulated telegraph and telephone.

Two wind-driven generators supplied power for transmitters, one being the main generator and the second a reserve one. In addition, batteries for emergency operation of the generators were supplied.

Three receivers of the superheterodyne type, designed to cover wavelength ranges of 150 to 500 meters, 400 to 4000 meters, and 3000 to 25000 were in operation.

A radio compass, with wavelength of 300 to 4000 meters, was extensively used. The compass coil was located in the buffer bag under the control car and was operated from the radio room.

Two antenna wires, approximately 650 feet in length, were suspended beneath the control car in which the radio equipment was located.

A short-wave transmitter and receiver for experimental work were also carried.

No data as to the effective operation of the radio equipment over long distances are available at this time. It is understood, however, that the *Zeppelin* was in communication constantly with land stations throughout the entire voyage.

Standard Frequency Transmissions by the Bureau of Standards

The Bureau of Standards announces its schedule of radio signals of standard frequencies for use by the public in calibrating frequency standards and transmitting and receiving apparatus. This schedule includes many of the border frequencies between services as set forth in the allocation of the International Radio Convention of Washington, which goes into effect January 1, 1929. The signals are transmitted from the Bureau's station WWV, Washington, D. C. They can be heard and utilized by stations equipped for continuous-wave reception at distances up to about 500 to 1,000 miles from the transmitting station.

The transmissions are by continuous-wave radiotelegraphy. The signals have a slight modulation of high pitch which aids in their identification. A complete frequency transmission includes a "general call" and "standard frequency" signal, and "announcements." The "general call" is given at the beginning of the 8-minute period and continues for about two minutes. This includes a statement of the frequency. The "standard frequency signal" is a series of very long dashes with the call letter (WWV) intervening. This signal continues for about four minutes. The "announcements" are on the same frequency as the "standard frequency signal" just transmitted and contain a statement of the frequency. An announcement of the next frequency to be transmitted is then given. There is then a 4-minute interval while the transmitting set is adjusted for the next frequency.

Information on how to receive and utilize the signals is given in Bureau of Standards Letter Circular No. 171, which may be obtained by applying to the Bureau of Standards, Washington, D. C. Even though only a few frequency points are received, persons can obtain as complete a frequency meter calibration as desired by the method of generator harmonics, information on which is given in the letter circular. The schedule of standard frequency signals is as follows:

RADIO SIGNAL TRANSMISSIONS OF STANDARD FREQUENCY SCHEDULE OF FREQUENCIES IN KILOCYCLES

Eastern Standard Time	Dec. 20	Jan. 21	Feb. 20	March 20
10:00–10:08 P.M.	4000	125	500	1500
10:12–10:20	4200	150	600	1700
10:24–10:32	4400	200	650	2250
10:36–10:44	4700	250	800	2750
10:48–10:56	5000	300	1000	2850
11:00–11:08	5500	375	1200	3200
11:12–11:20	5700	450	1400	3500
11:24–11:32	6000	550	1500	4000

Frequency Allocation Effective January 1, 1929

For the information of the membership of the Institute there is reproduced below the frequency allocation proposed by the International Radiotelegraph Conference held at Washington in 1927. This allocation will become effective throughout the world on January 1, 1929.

Members interested in a detailed account of the activities of the conference are referred to "International Radiotelegraph Conference", and "General and Supplementary Regulations Relating Thereto," which can be obtained from the Government Printing Office at Washington, D. C., for forty cents.

FREQUENCY ALLOCATION

Frequencies in kilocycles per second	Approximate wavelengths in meters	Services
10– 100	30,000–3,000	Fixed services.
100– 110	3,000–2,725	Fixed services and mobile services.
110– 125	2,725–2,400	Mobile services.
125– 150[1]	2,400–2,000[1]	Maritime mobile services *open to public correspondence exclusively*.
150– 160	2,000–1,875	Mobile services.
160– 194	1,875–1,550	(*a*) Broadcasting. (*b*) Fixed services. (*c*) Mobile services. The conditions for use of this band are subject to the following regional arrangements: All regions where broadcasting stations now exist working on frequencies below 300 kc (above 1000m). } Broadcasting Other regions { Fixed services. / Mobile Services Regional arrangements will respect the rights of other regions in this band.
194– 285	1,550–1,050	(*a*) Mobile services. (*b*) Fixed service. (*c*) Broadcasting. The conditions for use of this band are subject to the following regional arrangements: Europe { (*a*) *Air* mobile service *exclusively*. (*b*) *Air* fixed services *exclusively*. (*c*) Within the band 250–285 kc (1200–1050m). Fixed service *not open to public correspondence*. (*d*) Broadcasting within the band 194–224 kc (1550–1340m). Other regions { (*a*) Mobile services except commercial ship stations. (*b*) Fixed *air* services *exclusively*. (*c*) Fixed services *not open to public correspondence*.

[1] The wave of 143 kc (2,100m) is the calling wave for mobile stations using long continuous waves.

[2] The wave of 333 kc (900m) is the international calling wave for air services.

Frequency Allocations (*Cont.*)

Frequencies in kilocycles per second	Approximate wavelengths in meters	Services
285– 315	1,050 950	Radio beacons.
315– 350[2]	950– 850[2]	*Air* mobile services *exclusively*.
350– 360	850– 830	Mobile services *not open to public correspondence*.
360– 390	830– 770	(*a*) Radio compass service. (*b*) Mobile services, on condition that they do not interfere with radio compass service.
390– 460	770– 650	Mobile services.
460– 485	650– 620	Mobile services (*except damped waves and radiotelephony*).
485– 515[3]	620– 580[3]	Mobile services (distress, call, etc.).
515– 550	580– 545	Mobile services *not open to public correspondence* (*except damped waves and radiotelephony*).
550– 1,300[4]	545– 230[4]	Broadcasting.
1,300– 1,500	230– 200	(*a*) Broadcasting. (*b*) Maritime mobile services, waves of 1365 kc (220m) exclusively.
1,500– 1,715	200– 175	Mobile services.
1,715– 2,000	175– 150	Mobile services. Fixed services. Amateurs.
2,000– 2,250	150– 133	Mobile services and fixed services.
2,250– 2,750	133– 109	Mobile services.
2,750– 2,850	109– 105	Fixed services.
2,850– 3,500	105– 85	Mobile services and fixed services.
3,500– 4,000	85– 75	Mobile services. Fixed services. Amateurs.
4,000– 5,500	75– 54	Mobile services and fixed services.
5,500– 5,700	54– 52.7	Mobile services.
5,700– 6,000	52.7– 50	Fixed services.
6,000– 6,150	50– 48.8	Broadcasting.
6,150– 6,675	48.8– 45	Mobile services.
6,675– 7,000	45–42.8	Fixed services.
7,000– 7,300	42.8–41	Amateurs.
7,300– 8,200	41–36.6	Fixed services.
8,200– 8,550	36.6–35.1	Mobile services.
8,550– 8,900	35.1–33.7	Mobile services and fixed services.
8,900– 9,500	33.7–31.6	Fixed services.
9,500– 9,600	31.6–31.2	Broadcasting.
9,600–11,000	31.2–27.3	Fixed services.
11,000–11,400	27.3–26.3	Mobile services.
11,400–11,700	26.3–25.6	Fixed services.
11,700–11,900	25.6–25.2	Broadcasting.
11,900–12,300	25.2–24.4	Fixed services.
12,300–12,825	24.4–23.4	Mobile services.
12,825–13,350	23.4–22.4	Mobile services and fixed services.
13,350–14,000	22.4–21.4	Fixed services.
14,000–14,400	21.4–20.8	Amateurs.
14,400–15,100	20.8–19.85	Fixed services.
15,100–15,350	19.85–19.55	Broadcasting.
15,350–16,400	19.55–18.3	Fixed services.
16,400–17,100	18.3 –17.5	Mobile services.
17,100–17,750	17.5 –16.9	Mobile services and fixed services.
17,750–17,800	16.9 –16.85	Broadcasting.
17,800–21,450	16.85–14	Fixed services.
21,450–21,550	14 –13.9	Broadcasting.
21,550–22,300	13.9 –13.45	Mobile services.
22,300–23,000	13.45–13.1	Mobile services and fixed services.
23,000–28,000	13.1 –10.7	Not reserved.
28,000–30,000	10.7 –10	Amateurs and experimental.
30,000–56,000	10 – 5.35	Not reserved.
56,000–60,000	5.35–5	Amateurs and experimental.
Above 60,000	Below 5	Not reserved.

[3] The wave of 500 kc (600m) is the international calling and distress wave. It may be used for other purposes on condition that it will not interfere with call signals and distress signals.

[4] Mobile services may use the band 550 to 1,300 kc (545–230m) on condition that this will not cause interference with the services of a country which uses this band exclusively for broadcasting.

Note.—It is recognized that short waves (frequencies from 6,000 to 23,000 kc approximately—wavelengths from 50 to 13m approximately) are very efficient for long distance communications. It is recommended that as a general rule this band of waves be reserved for this purpose, in services between fixed points.

Institute Meetings

BUFFALO-NIAGARA SECTION

The September 17th meeting of the Buffalo-Niagara Section was held in Foster Hall, University of Buffalo. L. C. F. Horle, Chairman of the Section, presided.

E. T. Dickey and F. H. Engel, of the Radio Corporation of America, presented papers on "Quantitative Measurements Used in Tests of Broadcast Radio Receivers" and "Vacuum-Tube Production Tests," respectively.

Forty-three members of the Section attended the meeting.

On October 12th a joint meeting of the Section with the Niagara Frontier Section of the A.I.E.E. and the Radio Association of Western New York was held in Edmond Hayes Hall, University of Buffalo.

Two hundred and twenty members of the combined societies and their guests attended the meeting which was addressed by Richard H. Ranger, of the Radio Corporation of America, on "Recent Developments in Photoradio."

Mr. Ranger's address was supplemented with a demonstration of some of the photoradio devices described.

The next meeting of the Buffalo-Niagara Section will be held on November 14th in Foster Hall, University of Buffalo. I. R. Lounsberry and J. Morrison will be the speakers of the evening.

DETROIT SECTION

On September 20th a meeting of the Detroit Section was held in the Conference Room of the Detroit News Building. Earle D. Glatzel, Chairman of the Section, presided.

E. T. Dickey and F. H. Engel presented papers on "Quantitative Measurements Used in Tests of Broadcast Radio Receivers" and Vacuum-Tube Production Tests," respectively.

Thirty-six members of the Section attended the meeting.

Following the presentation of the papers, Messrs. Buchanan, Glatzel, Holland, and others participated in its discussion.

A meeting of the Detroit Section was held on October 19th in the Detroit News Building to hear a paper by Dr. H. B. Vincent, of the Engineering Research Department, University of Michigan.

Messrs. Holland, Katzin, Swain, and Glatzel, among others, participated in the discussion which followed the presentation of the paper.

The paper contained a discussion of the theory of detection of a triode for unmodulated and modulated signals of small magnitude, the developments described being taken from works published on the subject.

Los Angeles Section

The Los Angeles Section held a meeting on October 15th in the Elite Cafe, Los Angeles. Mr. Thomas McDonough, Vice-Chairman of the Section, presided.

B. F. McNamee read a paper "The Vacuum Tube As Applied to Radio Receiving Apparatus."

L. B. Park delivered a paper on "Telephoto and Transatlantic Telephony."

Forty-eight members of the Section attended the meeting.

New York Meeting

The New York meeting of the Institute for November was held on the 7th of the month in the Engineering Societies Building, 33 West 39th Street, New York City. Alfred N. Goldsmith, President of the Institute, presided.

Austin Bailey, of the American Telephone and Telegraph Company, presented a paper by Messrs. Bailey, Dean and Wintringham on "The Receiving System for Long-Wave Transatlantic Radio Telephony." The paper is printed elsewhere in this issue of the Proceedings.

In the discussion which followed the presentation of the paper Messrs. Van Dyke, Pratt, Bailey, and Wintringham participated.

Pittsburgh Section

The Pittsburgh Section held a meeting on September 18th in the Assembly Room of the Ft. Pitt Hotel, Pittsburgh. W. K. Thomas, Chairman of the Section, presided.

J. C. McKinley delivered a paper on "Elimination of Interference from Radio Receivers." The paper described the various methods which are used for the elimination of interference from radio receivers and included various sources of radio interference, which were placed in six general classes. These are static, heterodyning of stations and station noise, sparking applications, defective insulation, industrial applications, and radio receivers. The use of direction antenna in the elimination of interference was described. It was brought out that the underground antenna is selective and seems to be more sensitive to a definite frequency but the general results obtained showed no gain through its use.

Messrs. R. S. Johnson, L. A. Terven, Wolfe, Thomas, and Froelich participated in the discussion. Twenty members of the Section attended this meeting.

On October 9th a meeting of the Pittsburgh Section was held in the Norse Room of the Fort Pitt Hotel. L. A. Terven presided.

Austin Bailey, of the American Telephone and Telegraph Company, presented a paper on "The Receiving System for Long-Wave Transatlantic Radio Telephony." The paper is printed elsewhere in this issue of the PROCEEDINGS.

Messrs. Terven, Wolfe, Horne, McKinley, and Beers participated in the discussion which following the presentation of the paper.

Twenty-eight members of the Section attended the meeting.

SAN FRANCISCO SECTION

The San Francisco Section held its October meeting on the 24th of the month at the Bellevue Hotel, San Francisco.

Ralph M. Heintz, of Heintz and Kaufman, delivered an address on "Short-Wave Transmitting and Receiving Apparatus Designed Especially for Aircraft Service." Replicas of the radio equipment used on the Wilkins North Pole Expedition, the Byrd South Pole Expedition, and the aeroplane *Southern Cross*, which recently flew from San Francisco to Australia, were shown to the twenty-three members and seventeen guests who attended the meeting.

SEATTLE SECTION

On October 26th a meeting of the Seattle Section was held in the Club Room of the Telephone Building. Walter A. Kleist, Chairman of the Section, presided.

A paper on "Dynamic Exponential Loud Speakers" was delivered by Abner R. Willson, Walter A. Kleist, and J. R. Tolmie.

The paper included a series of brief notes on papers previously published by Hanna and others, followed by demonstrations by Messrs. Kleist and Tolmie.

Equations on design of the early horn type speakers followed by the balanced armature and cone types were set up and detailed design data on the plunger type diaphragm receiver unit working into an exponential horn was presented. Both the Hanna and the Wente and Thuras drive units were discussed. The demonstration included the operation of an ingenious switching system, developed by Mr. Kleist, by means of which the output of a new turn table and amplifier could be put through any one of six loudspeakers with equal power input. The early types of magnetic unit, the later improved types of magnetic unit, and the latest dynamic type of unit were all fitted with six foot exponential horns. The demonstration readily brought out the increase in volume and quality as the output of the amplifier was advanced from the older to the newer loudspeaker units.

Messrs. Libby, Deardorff, and Hackett participated in the discussion which followed.

Sixty-six members and guests attended the meeting.

The next meeting of the Seattle Section will be held on November 30th in the Club Room of the Telephone Building at Seattle. Professor Austin V. Eastman, of Washington University, will deliver an address on "The Four-Electrode Vacuum Tube."

WASHINGTON SECTION

On October 11th a meeting of the Washington Section was held in the Continental Hotel, 1 Capitol Street, Washington, D. C. F. P. Guthrie, Chairman of the Section, presided.

Austin Bailey, of the American Telephone and Telegraph Company, presented a paper on "The Receiving System for Long-Wave Transatlantic Radio Telephony." The paper is published elsewhere in this issue.

A. Hoyt Taylor, August Hund, H. G. Dorsey, G. D. Robinson, D. G. Howard, F. P. Guthrie, and E. B. Dallin participated in the discussion of the paper.

Forty-nine members of the Section attended the informal dinner preceding the meeting. Seventy-two members and guests attended the meeting.

Committee Work

COMMITTEE ON STANDARDIZATION

The meeting of the Committee on Standardization was held at the office of the Institute on October 16th. The following persons were present: L. E. Whittemore, Chairman; E. T. Dickey, E. N. Dingley, Jr., (representing Lieut. Com. W. J. Ruble), Dr. C. B. Jolliffe, Capt. Edwin R. Petzing (representing Maj. William R. Blair), E. L. Nelson, V. D. Landon (representing M. C. Batsel), Irving Wolff, (representing Subcommittee on Electro-Acoustic Devices), H. M. Turner, L. G. Bostwick (representing H. A. Frederick, of Subcommittee on Electro-Acoustic Devices), and E. W. Bemis, (representing Mr. H. S. Osborne).

The Committee continued its discussion of the reports of the various Subcommittees. The Committee will continue to meet at intervals of approximately two weeks until the final report has been completed for submission to the Board of Direction for formal adoption.

The final report of the Committee on Standardization, it is anticipated, will be published in the 1929 Year Book which will be mailed as a supplement to one of the issues of the PROCEEDINGS.

COMMITTEE ON ADMISSIONS

A meeting of the Committee on Admissions was held in the office of the Institute at 9:30 A.M. on November 7th. The following Committee members were present: R. A. Heising, Chairman; H. F. Dart, and E. R. Shute. The Committee considered twelve applications for transfer or election to the higher grades of membership in the Institute.

COMMITTEE ON MEMBERSHIP

A meeting of the Committee on Membership was held at dinner at the Fraternities Club Grill on November 7th. H. F. Dart, Chairman; H. B. Coxhead, and F. R. Brick were present.

The Committee considered a number of questions relating to means for increasing the membership of the Institute. It was announced that the net increase in members during the past year was over 600.

OBITUARY

With deep regret the Institute announces the death of

Charles Allen Wright

Mr. Wright was born in Vicksburg, Mississippi, in 1884. He graduated from Tulane University and received the M.E.E. degree from Harvard University in 1910. For five years he served as Professor of Electrical Engineering at Iowa State College, Carnegie Institute of Technology, and Ohio State University. In 1926 he was called to take charge of the Radio Division of the National Carbon Company Research Laboratories at Cleveland, a position he held up to the time of his death. He made numerous contributions to the technical literature and took active part in the meetings, discussions and Committee work of the various technical societies of which he was a member, among them this Institute, American Institute of Electrical Engineers, and Society for the Promotion of Engineering Education. At the time of his death he was a reserve officer in the Signal Corps of the U. S. Army with the rank of Captain.

His death will be mourned by a host of friends, associates, and former students.

Volume 16, Number 12 *December, 1928*

NOTE ON THE EFFECT OF REFLECTION BY THE MICROPHONE IN SOUND MEASUREMENTS*

By

Stuart Ballantine

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Summary—*The reflection of sound by the diaphragm of the microphone ordinarily used in the measurement of the instantaneous pressure in sound waves causes the indicated pressure to vary from equality with the actual pressure in the undisturbed wave at low frequencies, to twice this pressure at high frequencies. Due to the mathematically irregular shape of the conventional microphone and its mounting the effect cannot be calculated although it can be experimentally determined by comparison with the Rayleigh disk. It is proposed to evaluate the correction for reflection by employing a standard spherical mounting of which the diaphragm occupies a small area about the pole; the increase in pressure for this mounting can be calculated theoretically, and the correction for other mountings can then be obtained by experimental comparison.*

The ratio of pressure in the pole of the spherical mounting to that in the undisturbed wave, for a plane wave of the type Exp $i\omega(t-x/V)$ *is expressed in terms of Hankel's* $H^2_{n+1/2}$ *functions, for which tables exist up to the highest orders required for the computations in practical cases. Numerical computations are carried out in full, giving the vector pressure ratio at the pole facing the source for spheres of various diameters and at various frequencies throughout the acoustic range.*

In measuring or recording instantaneous sound-pressure variations with a calibrated condenser microphone it is often assumed that the pressure at the surface of the diaphragm is the same as that which would exist in the undisturbed sound wave; also some investigators have assumed that the pressure is *doubled* by reflection, therefore that the apparent values are to be divided by a factor of 2. If the diaphragm were of infinite extent, or part of an infinite wall, the pressure would clearly be doubled at all frequencies since the reflection coefficient at the air-membrane interface is very closely equal to unity due to the stiffness of the tightly stretched membrane. If, on the other hand, the dimensions of the microphone were small in comparison with the wavelength of sound, we should then have an ordinary problem of the Laplacian flow of air past an irregular obstacle and the pressure at the diaphragm would approach that in the undisturbed sound field. The effect of diffraction around the microphone then is to cause the apparent pressure (the pressure acting upon the diaphragm) to vary from the true pressure in the undisturbed wave, to twice this pressure as the frequency is raised from a low to a high value.

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* Original Manuscript Received by the Institute, September 21, 1928.

An obvious method of evaluating this effect and rendering useful the ordinary calibrations of the microphone derived from the application to the diaphragm of known alternating pressures produced by the thermophone, piston-phone or by electrostatic means, would consist in the direct comparison with the Rayleigh disk. The sound field at the point to be occupied by the microphone is first measured by means of the Rayleigh disk, the microphone is then substituted, and the overall calibration directly obtained in this way will include the effect of reflection. In the previous comparisons between the thermophone calibration of a condenser microphone and the Rayleigh disk calibration certain discrepancies have been noticed which are probably largely due to diffraction around the microphone.

In a paper to be published elsewhere[1] I have considered the mathematical theory of the effect and suggested a method for evaluating the appropriate corrections by means of a spherical mounting which can be calculated. In view of the increasing interest of radio engineers in acoustic matters I have ventured in the present note to summarize briefly the practical results of computations from this theory.

1. Standard Spherical Mounting. The geometrical volume occupied by the condenser transmitter, and single stage amplifier which is usually mounted with it for the purpose of avoiding long high capacity leads, is of a mathematically irregular shape and not amenable to calculation. To facilitate mathematical investigation we may mount the transmitter and amplifier in a rigid spherical shell with the diaphragm as nearly in the surface as possible. (Fig. 1). The diffraction of sound by a spherical obstacle is a classical problem,[2] and tables now exist which greatly facilitate the actual numerical computations.

When the relation between the pressure-ratio (actual pressure ÷ pressure in the wave in the absence of the obstacle) and frequency has been calculated for the standard spherical mounting, the effect of reflection with the more usual mountings may be readily evaluated by experimental comparison.

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[1] *Phys. Rev.*, Dec., 1928.

[2] Rayleigh: "Theory of Sound," 2, 218 et seq., (London, 1878): Papers, No. 287, V, p. 112, 1903; No. 292, V, 149, 1904.
Lamb: "Hydrodynamics," 5th Ed., p. 496, (Cambridge, 1924).

2. Diffraction of a Plane Sound Wave by a Rigid Sphere. Let the incident plane wave of sound be of the form $Exp\ i\omega(t-x/V)$, where V is the velocity of sound in air. The effect of diffraction may be represented by the ratio of the pressure (instantaneous) at the pole of the sphere facing the source to the pressure which would exist at the same point of space if the sphere were absent. Computation shows that the pressure at the pole is not substantially different from the average pressure over a small area about the pole so that in the practical case of a finite diaphragm area a negligible error is committed by using the polar pressure. The result of the theoretical calculation of the pressure ratio may be expressed as follows:

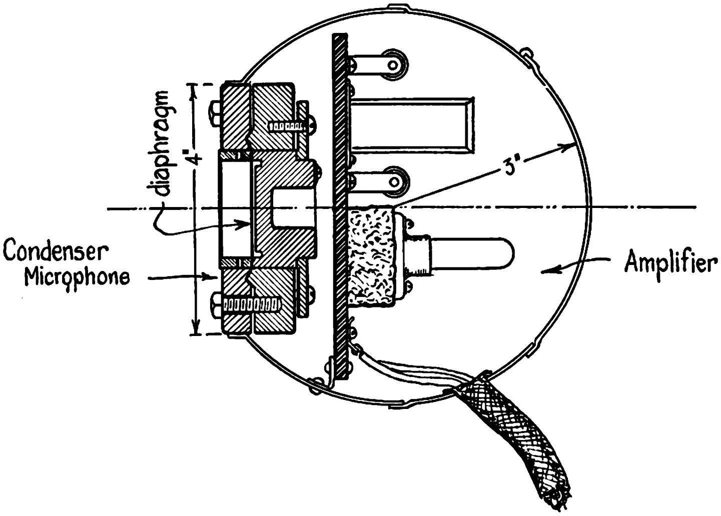

Fig. 1—Standard Spherical Mounting for Condenser Microphone.

$$\frac{P}{P_0}=e^{-ika}\sqrt{\frac{2}{\pi ka}}\sum_{n=0}^{\infty}\frac{-(2n+1)i^n}{(-)^n(nJ_{-n-1/2}+kaJ_{-n-3/2})-i(nJ_{n+1/2}-kaJ_{n+3/2})}. \quad (1)$$

(Argument of Bessel functions $=ka$)

The details of the derivation of this formula may be found in the more extensive paper.

3. Computations and Curves. The pressure ratio is a function only of the ratio of the size of the sphere to the wavelength of the sound. For computations from (1) tables are available[3] of $J_{n+1/2}$ for n up to 18, and of $J_{-n-1/2}$ for n up to 6; additional values of $J_{-n-1/2}$ for larger n may be computed from the tables of the related C_n functions given in the *British Association Reports* for 1914 and 1916.[4] Computations made with the aid of these modern tables are given in the annexed table:

TABLE I

Values of the vector ratio ($|p/p_0|\ expi\ \psi$) of pressure at pole of rigid sphere to pressure in incident plane wave at same point for various values of $ka = 2\pi \times$ *frequency* $\times$ *radius of sphere* $\div$ *velocity of sound in air.*

ka	p Equation (1)	$\left\lvert\frac{p}{p_o}\right\rvert$	Angle Ψ (radians)
0.1		1.0005	
0.2		1.0027	
0.3	$0.909+0.458i$	1.019	0.168
0.5	$1.043+0.299i$	*1.089	0.279
0.7	$0.597+1.040i$	1.198	0.345
0.85	$0.468+0.374i$	1.325	0.359
1.0	$0.310+1.370i$	1.406	0.340
2.0	$1.593+0.468i$	*1.660	0.286
3.0	$-1.762-0.165i$	1.772	0.235
4.0	$-0.910-1.595i$	1.835	0.194
6.0	$1.895-0.292i$	1.913	0.128
10.0	$1.540+1.207i$	*1.958	0.665 (?)

The angle ψ of the vector ratio is given as well as the absolute value. This is of interest in estimating the dispersion correction

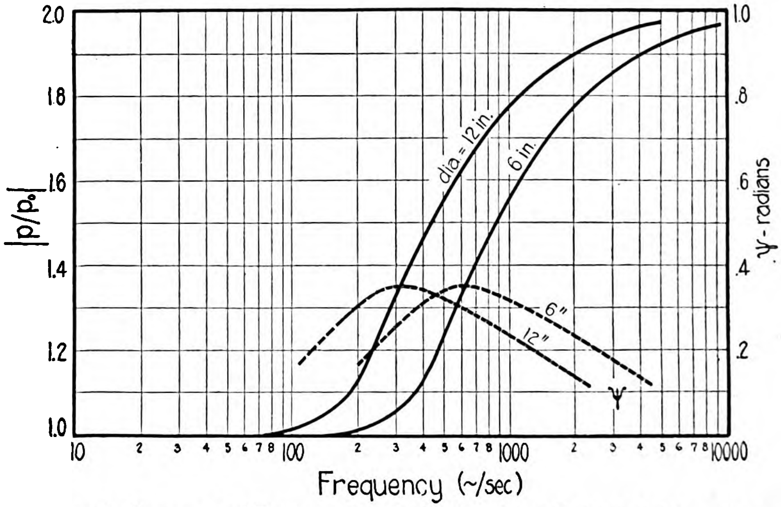

Fig. 2—Increase of Pressure by Reflection from Spherical Mounting. Absolute Value and Angle of Ratio of Pressure at Pole of Sphere to Pressure in Undisturbed Sound Wave.

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[3] G. N. Watson: "Bessel Functions," pp. 740–743 (Cambridge 1922).
[4] *British Association Reports*: p. 88–102, 1914; p. 97–107, 1916.

which must be made by means of phase equalizers in the electric circuits when an exact recording of the wave forms of the sound is desired. For many purposes, however, the absolute value of the ratio alone will be sufficient.

Fig. 2 contains a set of curves based upon these computations, which gives the diffraction correction for spherical mountings of various sizes over the audio range of frequencies. The form of the correction curves for other practical mountings may now be obtained experimentally by comparison with the standard mounting. This experimental program is under way and will be reported in a subsequent paper. It is necessary to investigate the functional relation for a single size of each shape; the position of the curve for any other size may be determined by the principle of similitude.

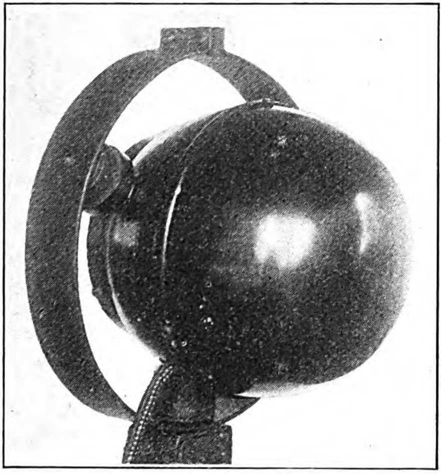

Fig. 3—Photographic View of Experimental Spherical Mounting.

A photographic view of the experimental spherical mounting is shown in Fig. 3. In the case where the condenser transmitter is mounted integrally with the apparatus comprising the first amplifier stage and occupies a volume of reasonable (say cubical) shape, it is often sufficient to estimate an "equivalent sphere" on the basis of equal volume to represent the irregular actual volume. The accuracy of this seemingly crude assumption is somewhat surprising.

Although the condenser microphone has been mentioned particularly in this discussion, it is obvious that it applies equally well to other pick-up instruments of the exposed diaphragm type, such as for example, the double carbon button microphone used extensively in radio broadcasting and public address systems. When the microphone is used for technical purposes and sound-wave recording it is convenient to compensate the diffraction effect either in the design of the microphone (which is feasible with air-damped types) or by equalization in the electrical circuits.

In conclusion it may be noted that by the reciprocity theorem equation (1) is equally appropriate for the representation of the pressure at a large distance in the sound radiated by a small piston located in the surface of a sphere. This throws some light on the action of baffles for loudspeakers.

Volume 16, Number 12 *December, 1928*

THE RECEIVING SYSTEM FOR LONG-WAVE TRANSATLANTIC RADIO TELEPHONY*

By

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Summary—*Transmission considerations and practical limitations indicate that in the lower frequency range, frequencies near 60 kc are best suited for transatlantic radio-telephone transmission. A radio receiving location in Maine gives a signal-to-noise ratio improvement over a New York location equivalent to increasing the power of the British transmitter about 50 times.*

Various types of receiving antennas are briefly discussed. The wave-antenna is selected as being most suitable for long-wave radio telephony. The various factors affecting wave-antenna performance and methods for measuring the physical constants of wave-antennas are discussed in detail. High-frequency ground conductivities determined from wave-antenna measurements are given. Combination of several antennas to form arrays is found to be a desirable means of decreasing interference. The use of a wave-antenna array in Maine decreases the received noise power by an additional 400 times. If the receiving were to be accomplished near New York using a loop antenna, we would have to increase the power of the British transmitting station 20,000 times to obtain the same signal-to-noise ratio. Comparisons of calculated and observed directional diagrams of wave-antennas and wave-antenna arrays are presented and discussed.

The transmission considerations governing the design of a radio receiver for commercial telephone reception are outlined.

Mathematical discussions of the wave-antenna, antenna arrays, quasi-tilt angle, and probability of simultaneous occurrence of telegraph interference are given in the appendices.

EARLY in October, 1915, engineers of the Bell System stationed in Paris heard the words "good-night Shreeve" which had been transmitted from Arlington. That date then marks the inception of transatlantic radio-telephone receiving. The progress which has been made in the radio-telephone receiving art since these first experiments is demonstrated by contrasting the homodyne receiver and the non-directional antenna then used with the present commercial receiving system employing double-demodulation of single side-band signals and an extensive array of wave-antennas forming a highly directional

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* Original Manuscript Received by the Institute, September 13, 1928. Presented before meetings of the following Institute Sections: Pittsburgh, October 9, 1928; Philadelphia, October 10, 1928; Washington, October 11, 1928; and at a New York meeting of the Institute, November 7, 1928.

system. In the pages which follow we shall endeavor to give some of the engineering considerations upon which the design of the present receiving system was based.

CHOICE OF FREQUENCY

In the early development of long-distance radio telegraphy, the strength of the received signal was the principal factor upon which the selection of the operating frequency was based. After the development of the vacuum-tube amplifier, however, the following considerations each became important, especially so for a telephone circuit:

1. The signal-to-noise ratio at the receiving location; which in turn is dependent upon four factors:
    - (a) The efficiency of the transmitting set
    - (b) The efficiency of the transmitting antenna
    - (c) Attenuation in the radio path
    - (d) Variation of radio noise with frequency
2. Band width of the transmitting antenna
3. Receiving antenna efficiency
4. Available space in the frequency spectrum.

1. *Signal-to-Noise Ratio at the Receiving Location.* At the time that the transatlantic radio-telephone development was undertaken engineers of the Western Electric Company Engineering Department (now Bell Telephone Laboratories) had developed a form of water-cooled vacuum tube capable of generating efficiently large amounts of power at any frequency up to perhaps several hundred kilocycles.[1] Therefore transmitter efficiency, although a major problem in itself, imposed no restriction on the frequency for the telephone circuit.

For transmission over a given path, utilizing a particular transmitting antenna with constant power supplied to it, there will be, in general, a frequency at which the greatest signal-to-noise ratio is obtained. To illustrate this point, we have chosen the problem of transmission from an antenna of the type used at the Rocky Point station of the Radio Corporation of America in U. S. A. to a receiving station in England, a distance of approximately 5,000 kilometers. The approximate variation with frequency of loss resistance, radiation resistance, and efficiency of this antenna is shown in Fig. 1. The loss resistance at 60 kilocycles was determined by engineers of Bell Telephone Laboratories, while the data in the lower frequency range were published by Alexanderson, Reoch, and Taylor.[2] The radiation resistance was calculated from the measured effective height of the antenna. It is seen in Fig. 1 that the antenna efficiency increases with frequency throughout the range we are considering, first rapidly and then more slowly.

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[1] W. Wilson, "A New Type of High Power Vacuum Tube," *Bell System Tech. Jour.*, 1, 4; July, 1922. *Elec. Comm.* 1, 15; August, 1922.

For a constant power radiated, radio attenuation tends to cause a decrease in the average received signal strength as the frequency is increased. This effect is in the opposite direction to the effect of antenna efficiency, so that for a given power supplied to the antenna the field strength at a given distance will be a maximum at a certain frequency. In Fig. 2 we have shown the calculated field strength at 5000 kilometers for a power of 85.9 kilowatts supplied to the Rocky Point antenna, using efficiency data of Fig. 1 and the radio transmission formula given by Espenschied, Anderson, and Bailey.[3] Since this curve reaches a maximum near 18.5 kilocycles, the reason for the operation of early transatlantic radio-telegraph circuits in the range 10 to 30 kilocycles becomes apparent in light of the limitation then placed on the receiving systems.

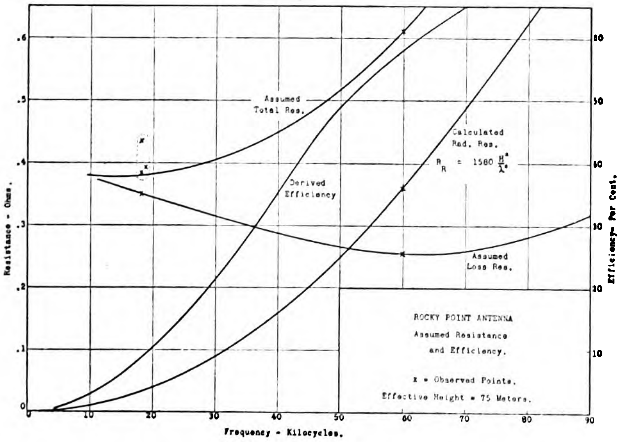

Fig. 1.

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[2] E. F. W. Alexanderson, A. E. Reoch, and C. H. Taylor, "The Electrical Plant of Transocean Radio Telegraphy," *Trans.* A.I.E.E., **42,** 707; July, 1923.

[3] Lloyd Espenschied, C. N. Anderson, and Austin Bailey, "Transatlantic Radio Telephone Transmission," *Bell System Tech. Jour.*, **4,** 459; July, 1925. Proc. I. R. E., **14,** 7; Feb., 1926.

Systematic measurements of radio noise by the warbler method,[4] begun early in 1923, have yielded important information on the variation of noise with frequency.[3] From measurements begun by engineers of Bell Telephone Laboratories and continued by engineers of the International Western Electric Company at New Southgate, England, during 1923 and 1924, the average daylight noise curve, in Fig. 2, was obtained. It is seen that the noise decreases with increasing frequency, at first rapidly and then more slowly, being almost constant after passing the frequency of 40 kilocycles.

From the values of signal and noise so obtained, the signal-to-noise ratio has been computed, and is also plotted in Fig. 2. The curve of signal-to-noise ratio reaches a maximum near 44 kilocycles, which would seem to be the optimum frequency for daylight transmission from the Rocky Point station to England. This is not strictly the case, however, since there is some evidence that a phenomenon exists which makes frequencies in the vicinity of 40 kilocycles particularly poor for the transatlantic path. Data published by Anderson[5] tend to show that the field strength is distinctly subnormal in the vicinity of 44 kilocycles and remains approximately constant from that frequency up to about 60 kilocycles, where the observed values agree fairly well with the calculations. (See later in this paper.)

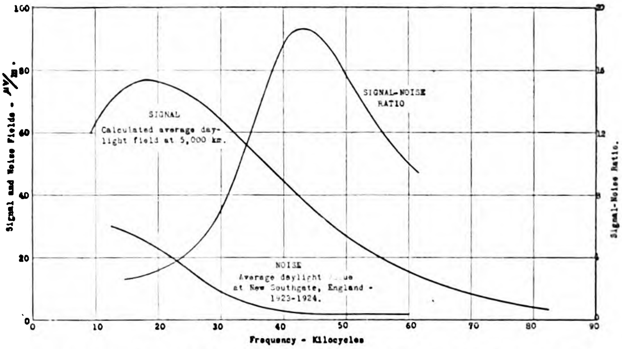

Fig. 2—Variation of Signal, Noise, and Signal-Noise Ratio with Frequency. Transmission from U. S. A. to England. 85.9 kw Supplied to Antenna of Rocky Point Characteristics.

[4] Ralph Bown, C. R. Englund, and H. T. Friis, "Radio Transmission Measurements," Proc. I. R. E., **11**, 115; April, 1923.

[5] C. N. Anderson, "Correlation of Long Wave Transatlantic Radio Transmission with Other Factors Affected by Solar Activity," Proc. I. R. E., **16**, 297; March, 1928. In connection with reference above see Fig. 19, p. 315.

2. Band-Width of the Transmitting Antenna. Since the output of the transmitting set is at a high power level, the circuits coupling it to the antenna must be of the simplest type to reduce the loss to a minimum. In view of this requirement, the antenna constants largely determine the band width of the antenna system. At frequencies much lower than 60 kilocycles it was not possible to secure a sufficient width of band even for commercial telephony from the Rocky Point antenna, but at this frequency reasonably satisfactory results are obtained.

3. Receiving Antenna Efficiency. The use of directional receiving antennas is essential to satisfactory and economic results over such distances as the transatlantic radio path (see later in this paper). The directivity of an antenna system of a given kind, size, and cost in general increases with frequency, since the directivity is a direct function of the ratio of the dimensions of the antenna system to the wavelength employed.

4. Available Space in the Frequency Spectrum. Each of the above factors operates to make the frequency of 60 kilocycles about the best which could be used in the present state of the art for this transmission path. Fortunately this frequency was so located in the radio spectrum that a band of the desired width free from interference could be obtained.

It has been noted that the radio noise as shown in Fig. 2 varies very little with frequency above 40 kilocycles. There is some doubt as to whether or not this accurately represents the actual state of affairs, since the measurement sets used for measuring the noise would not satisfactorily measure much below one microvolt per meter on account of tube noise. At frequencies of 40 kilocycles and above, especially in the winter, there are many days during which the radio noise is practically absent. On these days the measurements tended to approach the minimum determined by the set noise. The fact that many such readings were incorporated in the average probably tends to mask the true variations of radio noise with frequency in this range. On the other hand, however, they indicate a very real limitation which tends to operate against the use of frequencies higher than about 60 kilocycles unless fields were increased by increase in transmitting power. This would be particularly true during the sunset and sunrise dips and during periods of abnormally poor transmission when the fields fall much below the average. If the set noise limitation could be removed it is quite possible that frequencies above 60 kilocycles would become more useful. Higher frequencies for radio-telephone use would be particularly advantageous because of the greater band width which could be obtained from the transmitting antenna and because of the greater directivity which could be obtained in the receiving system at the same cost.

SELECTION OF A SATISFACTORY RECEIVING LOCATION

The selection of a suitable receiving location is based upon three major considerations; namely, maximum received signal-to-noise ratio, reasonably suitable terrain for receiving antenna construction, and adequate wire connection facilities between the the location and the more densely populated areas.

Since about 10 per cent of the populations of the United States and the British Isles are located within a radius of 40 miles of New York and London, respectively,[6] it was natural to decide upon making those cities the terminal points. It would hence be desirable to locate the receiving stations near and with good wire circuits to those cities.

Very early in the history of radio communication[7] it was, however, realized that in the United States a decrease of radio noise was obtained by a northerly location of the receiving station and, for receiving from European stations, the northern location is further advantageous, since higher field strengths result from the reduced transmission distance. The Radio Corporation of America had already taken advantage of this improvement by locating a receiving station at Belfast, Maine.

To obtain quantitative information on this matter the American Telephone and Telegraph Company made comparative measurements of noise as received on loop antennas at Riverhead, New York; Green Harbor, Massachusetts; and Belfast, Maine, the loops being so oriented as to give maximum receptivity in the direction of England. Although these tests were only continued for a few months at each location, they left no doubt that the absolute level of the noise was less at the northerly locations.

In Fig. 3, there is shown the diurnal variation of improvement in noise conditions (in TU) for average days of each month at Belfast over Riverhead. The average hourly improvement was determined by averaging the ratios of practically simultaneous observations of noise at the two locations for each hour during any one month and taking a three-hour moving average of the result to reduce the effect of purely local phenomena at either of the two stations. The data for the two half years were taken on slightly different frequencies as is indicated on the figure. Unfortunately, during the month of July only two weeks data were taken on each of the frequencies, namely, 52 and 65 kilocycles, and these data were taken a year apart, namely in 1924 and in 1925. In order to give some idea of the location noise improvement for the month of July we have averaged in the same way the four weeks data thus obtained, and plotted the result as a broken line. Fortunately, the improvement of the more northerly location is, in general, large during the overlapping business day of England and the United States.

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[6] "New York's New 10,000,000 Zone," *Literary Digest*, 95, No. 12, p. 14; Dec. 17, 1927.

[7] G. W. Pickard, "Static Elimination by Directional Reception," PROC. I. R. E., 8, 358; October, 1920.

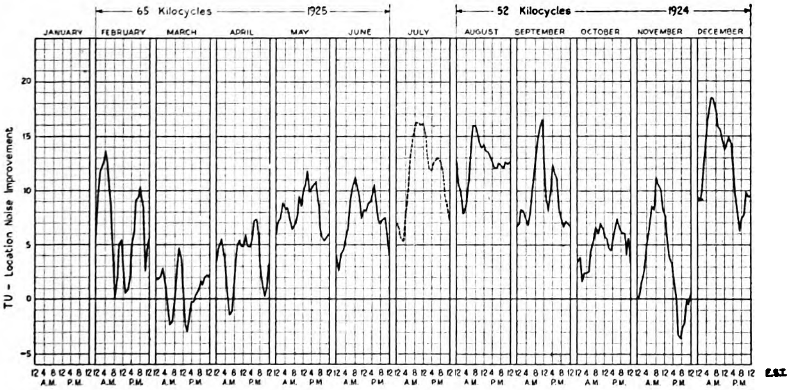

Fig. 3—Transatlantic Radio Noise Measurements. Diurnal Variations of Location Noise Improvement (in Transmission Units) of Belfast, Maine over Riverhead, New York. Three-Hour Moving Averages of Simultaneous Observations.

It is apparent that the improvement is a maximum in the middle of the summer when the noise is high, and in the middle of the winter when the field strengths are usually abnormally low. This is important, since the greatest improvement is needed at each of these times.

The monthly averages of variations of noise and of signal have previously been published[3,5,8], and the generalizations given above can be confirmed by reference to these articles.

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[8] Ralph Bown, "Some Recent Measurements on Transatlantic Radio Transmission," *Proc.* Natl. Acad. of Sci., 9, 221; July, 1923.

For calculating daylight radio transmission, several formulas have been proposed.[9,10,11] In Fig. 4 the heavy curve was calculated from the empirical formula given by Espenschied, Anderson and Bailey,[3] and assumes a radiated power of 50 kilowatts. The great circle distance from the transmitting stations used in the transatlantic radio-telephone circuit to various receiving stations is indicated by the name of the receiving station. The average of daily averages of hourly measurements of the field strength made at Houlton and Wroughton during the time that the transatlantic path was entirely in daylight during 1927 is indicated by points on this figure. The data for Cupar are less complete since this station was not in regular daily operation until May, 1927. The range of variation between the maximum daily average and the minimum daily average for each receiving location is given by the limits of the dotted vertical line. (It is interesting to note that at a frequency of 60 kilocycles and for distances in the

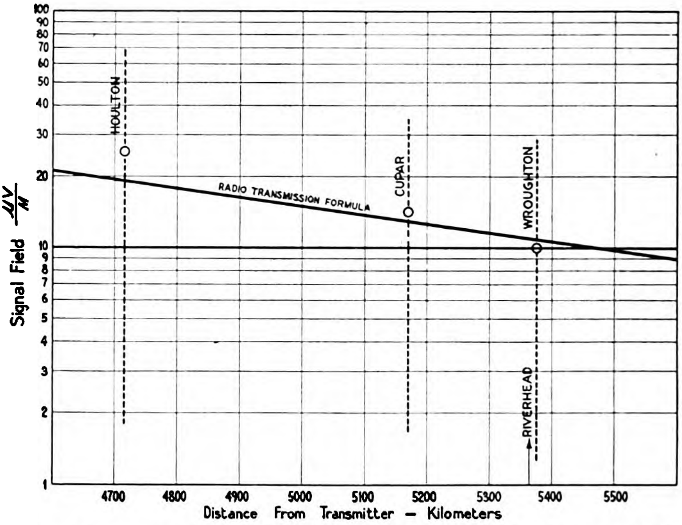

Fig. 4—Transatlantic Radio Daylight Field Strength. Average of Hourly Observations during 1927. Corrected to 50 kw Radiated Power—Frequency 60 kilocycles.

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[9] A. Sommerfeld, "Ueber die Ausbreitung der Wellen in der drahtlosen Telegraphie," *Ann. d. Phys.* **28,** 665; 1909.

[10] L. F. Fuller, "Continuous Waves in Long-Distance Radio Telegraphy," *Trans.* A.I.E.E., **34,** pt. 1, 809; 1915.

[11] L. W. Austin, "Quantitative Experiments in Radiotelegraphic Transmission," *Bull. Bureau of Std.,* **11,** 69; Nov. 15, 1914.

order of 5000 kilometers any of the radio-transmission formulas referred to above will give a computed value lying within the range of variation of average daylight readings.)

The improvement in signal-to-noise ratio obtained by locating the receiving station in Maine instead of in New York is easily seen by reference to Figs. 3 and 4. The improvement due to decrease of noise, during that time of year when improvements are most needed on account of high noise values, is about 10 TU. The improvement due to increase of the average received daylight signal by decrease of the distance is calculated to be 5 TU. During 1927, this improvement was actually observed to be 8 TU. We may, therefore, state in round numbers that the total improvement realized by locating the receiving station in Maine instead of New York was equivalent to a fifty-fold increase of the power radiated by the British transmitting station.

The British General Post Office, during 1926, carried out a set of measurements of field and noise at various locations in the United Kingdom. Those tests led them to the same conclusions as regards the advantage to be obtained by locating their receiving station at some more northerly point.[12] They decided upon a location near Cupar, Scotland, and comparisons made daily from 1230 to 2300 GMT indicate that this location is better for receiving than Wroughton, England. The geometric mean of the improvement in signal-to-noise ratio for the more northerly location during the months May to September, 1927, inclusive, and for the daily period given above is 6.4 TU. This is equivalent to an increase of between four and five times in power from the American transmitting station.

Since such relatively large improvements were to be obtained by northerly locations of the receiving station it seemed best to take advantage of this fact and locate the receiving station in America at some place in the state of Maine. This decision led to further consideration of two factors mentioned above, namely, reliable wire connections to New York and a suitable terrain for antenna construction. The first of these factors required a location along one of the main telephone trunk routes in Maine and the second, since we had decided upon the use of a wave-antenna[13] for reasons which will be given in the following section, demanded a rather large and reasonably flat land area available for pole-line construction. A location, although not altogether ideal, was decided upon near Houlton, Maine, about six miles from the Canadian border.

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[12] A. G. Lee, "Wireless Section: Chairman's Address, *Jour.* I.E.E. **66**, 12; Dec., 1927.

[13] H. H. Beverage, C. W. Rice and E. W. Kellog, "The Wave Antenna," *Trans.* A.I.E.E., **42**, 215; 1923.

CHOICE OF RECEIVING ANTENNA SYSTEMS

The number of fundamental types of receiving antennas that may be employed for long-wave reception is quite definitely limited. In fact all of the known practical receiving antennas may be considered as falling into one of three principal classes of structure; i.e., the vertical antenna, the loop or coil antenna, and the wave-antenna. The selection of the proper receiving antenna system quite evidently becomes a problem; first, of choosing the best type of antenna from one of these three classes and, second, of choosing a particular antenna structure in the class which is found to be best.

The factors governing the choice of a receiving antenna are as follows:

1. Directional Discrimination Against Static. Inasmuch as the signal to be received has a definite average value, the receiving system can only better the circuit in the amount that it improves the signal-to-noise ratio. A directional antenna system affords a means of reducing the received noise in relation to the desired signal.[7,13] The directional characteristics of the principal antenna types are shown in Fig. 5.

A measure of the directional discrimination of the various antenna types is the Noise Reception Factor (abbreviated *NRF*) which is defined as the ratio of the total noise current received from the antenna in question to that received from a vertical antenna under the conditions of continuous, constant distribution of noise sources about the antenna and of equal output currents for signals from the direction of maximum receptivity. The back end *NRF* is the noise reception factor for the arc between 90 degrees and 270 degrees from the direction of maximum receptivity.

On this basis, the choice rests quite unmistakably with the wave-antenna.

2. Transmission-Frequency Characteristic. Since the receiving antenna is to be used on a system for communication by speech, necessitating the transmission of a relatively wide band of frequencies, it must pass such a band without undue discrimination against any frequency contained therein. To utilize the vertical and the loop antennas efficiently, it is desirable that they be tuned, introducing the frequency discrimination of a tuned circuit. If these types of antennas be used it is necessary, therefore, that the resonance characteristic be studied and means provided to eliminate excessive frequency discrimination within the desired band. On the other hand, the wave-antenna is an aperiodic structure and, in consequence, its transmission-frequency characteristic is so flat that it need not be considered.

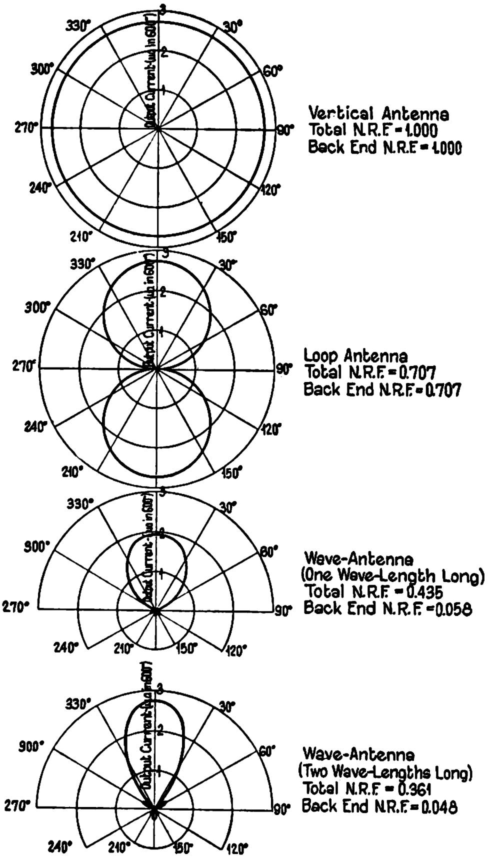

Fig. 5—Comparison of Simple Antennas.

3. Sensitivity. There are two factors which require that the output from the receiving antenna for a given field strength be as large as possible. First, if the receiving station be located at any position other than that at the terminal of the antenna, which is necessarily the case if more than one antenna be used in an array, the signal on the transmission line from the antenna to the station must be much greater than the noise currents induced into the transmission lines. If the antenna output be excessively small, it is impossible to balance the transmission lines so completely that this requirement is met. Second, the amount of gain that can possibly be used at the radio receiver is ultimately limited by the noise produced in an amplifier. (This is discussed more fully under "Power Output Required from the Radio Receiver" later in this paper.) To the first approximation, the sensitivity of each of the antenna classes under consideration is a direct function of its physical dimensions. There is, however, a limit to the sensitivity of each antenna class, for mechanical limits govern the maximum size of a vertical antenna, distributed capacity and mechanical considerations limit the loop, and in the wave-antenna a restriction occurs because of the peculiarity that the sensitivity reaches maximum values at definite lengths.

Since cost is likewise a factor governing the ultimate selection of an antenna system, the sensitivities may well be compared for antennas of equal cost. On this basis, a loop or a vertical antenna of effective height of fifty meters is directly comparable with a wave-antenna one wavelength long. By reference to Fig. 5, where the scale is the same for all the directional diagrams, it becomes evident that the sensitivities of all three classes of antennas are of the same order of magnitude, being slightly greater for the vertical antenna and the loop than for the one-wavelength wave-antenna.

4. Stability. The sensitivity and frequency-transmission characteristics of the antenna must be substantially constant during changes of weather and seasonal conditions. The antenna classes which require tuning are slightly poorer than the wave-antenna in this respect.

5. Reproducibility. Further improvement in directional discrimination against noise is obtained by using several similar antennas in an array. The loop and the vertical antennas probably are best for combining in arrays because several of either type of antenna can be made identical with one another. Wave-antennas combined in an array, however, give satisfactory results.

Although each of these factors governing the choice of the receiving antenna system is important, their relative impor-

tance is indicated by the order in which they have been presented. In view of the low noise-reception-factor of the wave-antenna, its lack of frequency discrimination, and its inherent stability, the wave-antenna was selected for the fundamental type of antenna to be used at the receiving station at Houlton.

THE WAVE-ANTENNA

Among the types of antennas which may be considered for use in long-wave radio communication, the wave-antenna[13] possesses several characteristics which single it out as being unique. The most important of these are:

1. The length of a wave-antenna is directly comparable to and of the same order of magnitude as the wavelength of the signals for which it is designed.
2. Considering the straight horizontal wire comprising the wave-antenna as a grounded transmission line, a termination, equal to the characteristic impedance, is applied to each end of that line The wave-antenna then becomes an essentially aperiodic antenna.
3. The major response of a properly designed wave-antenna is to the horizontal component of the impressed electric field. The propagated electric wave must therefore have an electric component parallel to the surface over which the wave-antenna is constructed.
4. On the basis of the preceding consideration, the design of a wave-antenna definitely excludes elevation of the antenna above ground to any extent greater (a) than is physically necessary to provide safe clearance and (b) than that height where the loss in the antenna considered as a transmission line reaches a nominal value. Practically, the wave-antenna is constructed as a high-grade telephone line, on 30-foot poles.

It is evident that the major electrical characteristics which distinguish the wave-antenna are intimately connected with the character of the surface over which the antenna is built, and with the details of construction of the wave-antenna. The performance of a wave-antenna at any specified location then can only be determined by constructing such an antenna and measuring its constants. The measurements made in determining the characteristics of any particular wave-antenna are outlined in the following paragraphs.

1. Ground-Connection Impedance. It is shown in Appendix 1 that the wave-antenna is considered to be a smooth line with uniformly distributed constants. This assumption is met to a sufficient degree in practice, but, unfortunately, it is impossible to connect to the four terminals of the practical line, since the connections to the ground side of the line must be made by burying wires in the earth rather than connecting to a discrete terminal which is the real ground. As is shown in Fig. 6a, the actual wave-antenna may still be considered as a smooth line, but between the terminals of the wave-antenna and the terminals that are available at the physical ends of the wave-antenna ground-connection impedances exist. To determine the constants of the wave-antenna, these impedances must be evaluated and taken into account as follows: In Fig. 6a, an impedance Z is applied to the available terminals of the wave-antenna 3–4 and the impedance S measured at the available terminals 1–2; under this condition, the actual terminal and input impedances of the wave-antenna are respectively:

$$Z' = Z + G_2 \qquad (1)$$

$$S' = S - G_1 \qquad (2)$$

where G_1 and G_2 are the ground-connection impedances at the two ends of the antenna.

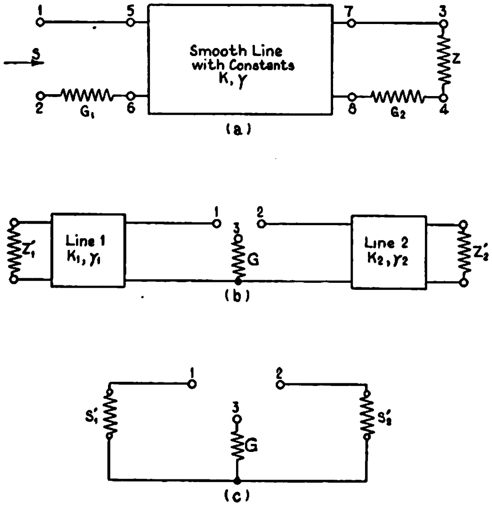

Fig. 6.

Figs. 6b and 6c illustrate the method that was used to determine the ground-connection impedance. In Fig. 6b, lines 1 and 2 represent two smooth ground-return transmission lines extending in opposite directions from the ground-connection for 1/2 kilometer or more, the lines being terminated at the distant end in impedances Z_1' and Z_2', respectively. In practice one of these lines was the wave-antenna and the other a temporary line of insulated wire laid along the surface of the ground.

For the purpose of analysis, each of the lines may be replaced by its input impedance. This simplification is shown in Fig. 6c, where

$$S' = \frac{K \tanh \gamma s + Z'}{1 + \frac{Z'}{K} \tanh \gamma s} \tag{3}$$

The impedance between terminals 1 and 3 is:

$$S_1 = S_1' + G \tag{4}$$

The impedance between terminals 2 and 3 is:

$$S_2 = S_2' + G \tag{5}$$

The impedance measured between terminals 1 and 2 in parallel and terminal 3 is:

$$S_0 = G + \frac{S_1' S_2'}{S_1' + S_2'} \tag{6}$$

Eliminating S_1' and S_2' from equations (4), (5), and (6) and solving for G:

$$G = S_0 - \sqrt{\frac{1}{2}[(S_1 - S_0)^2 + (S_2 - S_0)^2 - (S_1 - S_2)^2]} \tag{7}$$

By building out either line 1 or line 2 with added series impedances until

$$S_1 = S_2 = S_{12} \tag{8}$$

the expression for the ground-connection impedance simplifies greatly, and incidentally the precision of the determination becomes greater because the number of measurements involved is less. Under this condition

$$G = 2S_0 - S_{12} \tag{9}$$

This latter case is the one that was actually used in measuring the ground impedances.

Since the distribution of ground currents about the buried ground may be different under each of the three conditions that are measured, there is undoubtedly some error in measuring the ground-connection impedance by this method. This error is a second-order effect, however, so that the values determined are reliable within the precision that the method allows, involving as it does, differences between measurements of high-frequency impedance.

All of the impedance measurements were made using a high-frequency bridge designed and constructed by Mr. C. R. Englund of Bell Telephone Laboratories. This bridge is similar to that described by Shackelton[14] except that the standards used consist of a calibrated condenser and a decade resistance. Impedances having capacitive reactance are measured by direct comparison with the standards, while impedances having inductive reactance are tuned with the standard condenser to parallel resonance and the resonant combination compared with the decade resistance. Impedances involving extremely small reactances, either positive or negative, are built out with a condenser in parallel to a value that may be measured conveniently.

2. Characteristic Impedance and Propagation Constant. Since the early days of transmission line study, the characteristic impedance and the propagation constant have been determined by two impedance measurements at the near end of the line with the far end of the line open- and short-circuited, respectively.[15] For two reasons, this method has not been used in our determination of the fundamental antenna constants: first, it is impossible to apply a short to the real terminals of the wave-antenna due to the presence of the ground-connection impedance; and, second, with lines multiple quarter wavelengths long the input impedance, as a result of resonance in the line when it is open-circuited or grounded, attains either extremely large or extremely small values which could not be measured accurately with the available testing equipment.

To obviate these difficulties, Mr. C. R. Englund, of Bell Telephone Laboratories, developed a method of determining the characteristic impedance and the propagation constant of the wave-antenna by measuring the input impedance with two known finite terminations at the far end. Under this condition it may be shown that the characteristic impedance is given by the expression:

$$K=\sqrt{\frac{(S_1-G_1)(S_2-G_1)(Z_1-Z_2)+(Z_1+G_2)(Z_2+G_2)(S_2-S_1)}{(S_2-S_1)+(Z_1-Z_2)}} \qquad (10)$$

[14] W. J. Shackelton, "A Shielded Bridge for Inductive Impedance Measurements at Speech and Carrier Frequencies," *Bell System Tech. Jour.*, **6**, 142; Jan., 1927.

[15] Bela Gati, "On the Measurement of the Constants of Telephone Lines," *The Electrician*, **58**, 81; Nov. 2, 1906.

and that the propagation constant is given by:

$$\gamma = \frac{1}{s} \tanh^{-1} \left[K \frac{(S_2 - S_1) + (Z_1 - Z_2)}{(Z_1 + G_2)(S_1 - G_1) - (Z_2 + G_2)(S_2 - G_1)} \right] \qquad (11)$$

where the symbols in equations (10) and (11) have the following meanings:

Z_1 = the first termination applied to the available terminals at the far end of the line (ohms)
Z_2 = the second termination applied to the available terminals at the far end of the line (ohms)
S_1 = the impedance measured at the available near-end terminals corresponding to the termination Z_1 (ohms)
S_2 = the impedance measured at the available near-end terminals corresponding to the termination Z_2 (ohms)
G_1 = the ground-connection impedance at the near-end of the line (ohms)
G_2 = the ground-connection impedance at the far-end of the line (ohms)
s = length of the line (kilometers)
K = characteristic impedance (ohms)
γ = propagation constant (hyps per kilometer)

3. Effective Height. The effective height of a wave-antenna is defined as the ratio of the voltage produced at any specified point in the antenna to the potential gradient of the electromagnetic field producing that voltage. If the constants of the antenna system are known, the effective height at any point in the antenna system may be calculated from the value at any other point in the system.

A convenient way to measure an effective height of a wave-antenna and obtain a value which may be easily correlated with wave-antenna theory is to introduce in series with the initial-end terminating impedance a voltage which produces the same output current from the antenna as is produced by an electromagnetic wave. The ratio of this induced voltage to the potential gradient of the electromagnetic field has been called "the effective height referred to the characteristic impedance." For small values of the quasi-tilt angle, the total potential gradient of the electric field is very closely equal to the vertical component of the electric field, so that within the precision of measurement we may write:

$$H_e = \left| \frac{E_K}{E'} \right| \qquad (12)$$

where

H_e = Effective height of the wave-antenna referred to the characteristic impedance (kilometers)

E' = The potential gradient of the vertical component of the impressed field. (volts per kilometer)

E_K = The electromotive force introduced in series with the characteristic impedance at the initial end of the wave-antenna producing the same current at the distant end as the impressed field. (volts)

4. Quasi-tilt Angle and Ground Resistivity. The measured effective height of a wave-antenna is a function of four constants:

1. The length of the antenna.
2. The height of the antenna.
3. The propagation constant of the antenna.
4. The ratio of the component of the electric wave parallel to the surface over which the antenna is constructed to the vertical component of the electric wave.

In general, the first three of these constants are different in value for antennas constructed at different locations, but they may be varied over a limited range by changing the construction and dimensions of the wave-antenna. The comparison of effective heights, therefore, does not readily yield information regarding the relative suitability of various locations for wave-antenna systems. The ratio of the horizontal component to the vertical component of the impressed field is, however, a constant whose value is dependent solely upon the ground conditions at the location (assuming a fixed frequency for the comparison).

In case the time phase between the horizontal component and the vertical component of the impressed field were zero, the ratio of these two components would represent the tangent of the angle of forward inclination of the propagated wave front. In general, the phase angle between the two components is not zero, so that such a simple relation does not hold. It is convenient, however, to call the ratio of the two components of the impressed field the tangent of the "quasi-tilt angle", where the "quasi-tilt angle" becomes the real tilt angle in the limiting case.

In terms of the effective height, the antenna constants and the vertical component of the impressed field, the current produced at the far end of the wave-antenna is (using the nomenclature of Appendix 1 and to the same degree of approximation as equation (12)):

$$|I_\theta| = \left|\frac{H_\theta E' \epsilon^{-\gamma S\lambda'}}{2K}\right| \qquad (13)$$

or

$$|I_\theta| = H_\theta \epsilon^{-\alpha S\lambda'} \left|\frac{E'}{2K}\right| \qquad (14)$$

In terms of antenna constants alone, it is shown in Appendix 1 that the current produced at the far end of the wave-antenna is:

$$|I_\theta| = |I_{E'\theta} + I_{F'\theta}| \qquad (125)$$

where

$$I_{E'\theta} = -\frac{1}{\epsilon^{j\delta} \tan T} \frac{S\lambda' F}{2K}(a+jb) \qquad (15)$$

$$I_{F'\theta} = \frac{S\lambda' F'}{2K}(c+jd) \qquad (16)$$

also

$$-\frac{F'}{E'} = \epsilon^{j\delta} \tan T \qquad (301)$$

In equations (15) and (16), $(a+jb)$ and $(c+jd)$ are abbreviations defined as follows:

$$(a+jb) = \frac{h}{S\lambda'}\left(1 - \epsilon^{-[\alpha S\lambda' + j2\pi S(m - \cos\theta)]}\right)\epsilon^{-j2\pi S \cos\theta} \qquad (17)$$

and

$$(c+jd) = \cos\theta \, \frac{1 - \epsilon^{-[\alpha S\lambda' + j2\pi S(m - \cos\theta)]}}{\alpha S\lambda' + j2\pi S(m - \cos\theta)} \epsilon^{-j2\pi S \cos\theta} \qquad (18)$$

If we equate expressions (14) and (125), and solve for tan T:

$$\tan T = \frac{ac + bd + \sqrt{(ac+bd)^2 - (c^2+d^2)(a^2+b^2-g^2)}}{c^2+d^2} \qquad (19)$$

where

$$g = \frac{H_\theta}{S\lambda'} \epsilon^{-\alpha S\lambda'} \qquad (20)$$

It is pointed out in Appendix 3 that the phase angle δ may be expressed as a function of the quasi-tilt angle T and the dielectric constant k. For that reason, the determination of T must be made in two steps. The procedure is as follows: first, it is assumed that the component of the total received current due to the vertical component of the impressed field is zero, i.e.,

$$(a+jb)\equiv 0$$

Under this condition:

$$T=\tan^{-1}\frac{g}{\sqrt{c^2+d^2}} \tag{21}$$

Using Fig. 20 of Appendix 3, the value of δ corresponding to this value of T is determined (generally $\delta=\pi/4$). Second, T is revaluated from (19) using the value of δ so obtained.

The ground resistivity is evaluated from the value of the quasi-tilt angle by using Fig. 20 of Appendix 3.

5. *Directional Characteristics.* The measurement of the directional characteristics of a wave-antenna or a wave-antenna system consists entirely of measuring the effective height of the antenna for several directions of wave propagation, and determining the relative directional receptivity of the antenna in these directions by dividing the effective height for each direction by the value obtained for the direction of the axis of the antenna. For this purpose, the effective height at the output of the antenna system is most convenient to measure and use. This constant is defined as the ratio of the voltage at the input of the radio receiver to the field strength producing this voltage. It is exactly related to the effective height referred to the characteristic impedance (defined in the preceding subsection of this paper) by the real part of the total transfer constant between the termination at the initial end of the antenna and the input terminals of the radio receiver, and an additional factor of one-half because the voltage at the radio receiver is measured across the proper termination.

In certain receiving station locations, it is possible to utilize for determining the relative directional receptivity the regular transmission from existing radio transmitters operating at or very close to the frequency for which the directional characteristic is desired. At sites less favorably located with regard to existing transmitters, the directional characteristic may be measured by transmitting test signals from a portable transmitter, located successively in the several directions for which data are desired, and at least 15 wavelengths from the antenna system.

A distinctly different method of measuring the directional characteristics of an antenna is based on a statistical study of the reduction of noise obtained by its use. While it is difficult to evaluate the directional characteristic exactly by this method, data showing the comparative decrease in noise with the wave-antenna as against a loop or a vertical antenna are of great value in predicting the improvement in a radio circuit to be obtained by its use. As a converse to these results, the statistical combination of the improvement given by the wave-antenna, and a measured directional diagram, yields information on the direction of arrival of static.

Data on wave-antenna characteristics have been taken at several widely separated locations. Two antenna systems have been constructed by the British General Post Office, one at Wroughton, Wiltshire, in southern England, and one at Cupar, Fifeshire, in southeastern Scotland. We likewise have data on our antenna system at Houlton, Maine. The character of the earth under each of these antenna systems is different, resulting in widely different quasi-tilt angles and antenna directional characteristics.

The probable geological formations under individual antennas at each of the three antenna sites mentioned in the preceding paragraph are shown in Fig. 7. The data for the British locations were compiled from the published reports of geological surveys conducted by the British Government, and the data for the Houlton location were taken from the "Soil Survey of the Aroostook Area, Maine," published by the U. S. Department of Agriculture. In Table I, the ground constants are given for these three locations, determined by the method given in Section 4, "Quasi-tilt Angle and Ground Resistivity":

TABLE I

Location	Characteristic Sub-Soil	Quasi-Tilt Angle at 60 kc Radians	Ground Resistivity Ohms per cm^3
Wroughton	Chalk	0.011	3630.
Cupar	Sandstone	0.017	8670.
Houlton	Limestone	0.047	66300.

Fig. 8 shows the directional characteristics, calculated by the method given in Appendix 1, of wave-antennas erected over the geological formations shown in Fig. 7. In these directional characteristics, it is important to notice that a decrease in quasi-tilt angle increases the relative importance of the component of the received current due to the vertical component of the impressed field (abbreviated to the "vertical effect"). It is evident from Fig. 8 that the relative directional receptivity for the arc between $\Theta = 90$ degrees and $\Theta = 270$ degrees is much smaller and that the effective height is much greater for the antenna at Houlton, for which the ground resistivity is higher than for the other two antennas.

Measured relative directional receptivities are also shown in Fig. 8. The values for the Cupar antenna were determined by using transmission from the several European transmitting stations which are designated on this figure. The measurements on the Houlton antenna system were made using a portable two-kilowatt transmitter located successively at each of the 22 positions shown on the map, Fig. 9. The authors wish to thank Mr. G. D. Gillett for his efficient operation of this transmitter during the summer of 1927.

It is seen that the agreement between the measured and the computed directional characteristic is much better for the shortened Houlton *A* antenna than it is for the same antenna 0.70 kilometer longer. The reason for this can be appreciated by reference to Fig. 7, where it is shown that the far end of the long antenna is at the top of a rocky hill, while after shortening the far end is in a swamp, at the same average elevation as the remainder of the antenna. The elimination of this sharp rise over rocky ground serves principally to remove an irregularity in the constants of the wave-antenna near the end, so that the entire antenna may be considered more nearly a smooth line. This makes the antenna function more satisfactorily as a unit of an array in connection with other antennas constructed nearby.

6. Wave-Antenna Arrays. Since 1899, when S. G. Brown[16] proposed the use of two vertical antennas, separated in space by an appreciable portion of a wavelength and excited at a half-period phase difference, as a means of directional transmission, the use of arrays of antennas for directional transmission and reception has become increasingly important. Antenna arrays

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[16] R. M. Foster, "Directive Diagrams of Antenna Arrays," *Bell System Tech. Jour.*, 5, 292; April, 1926. Also see references listed in Foster's paper.

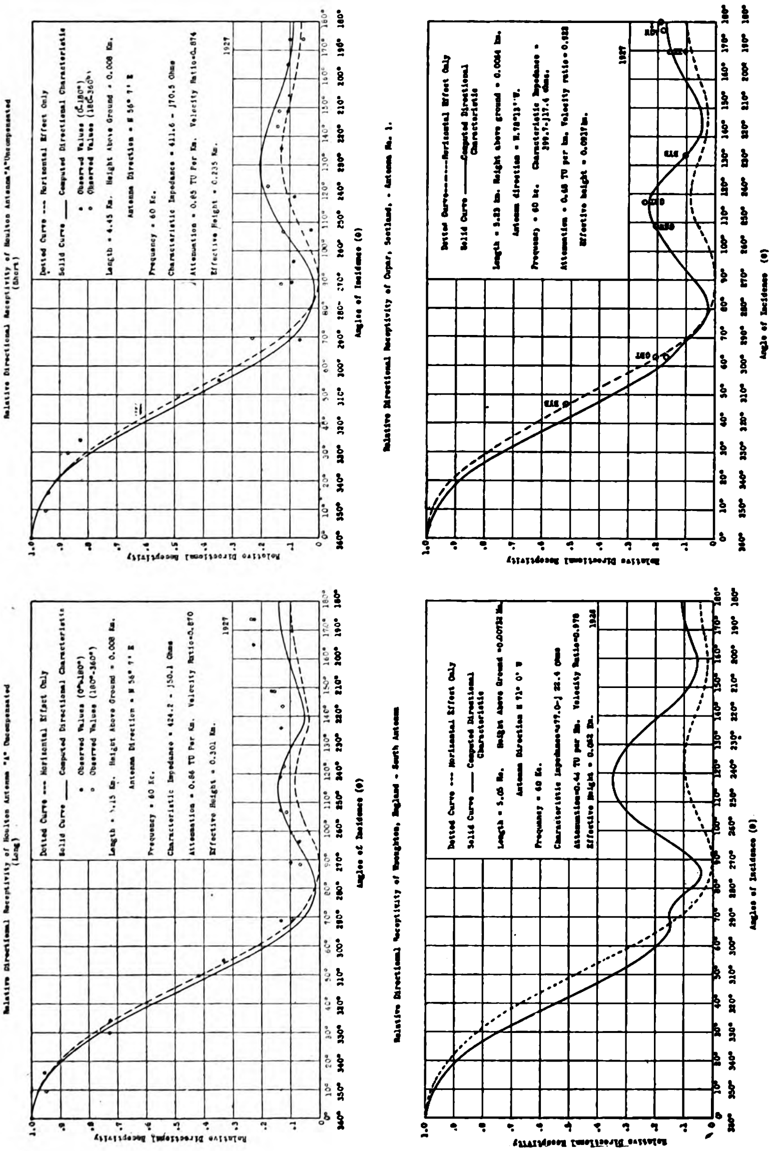

Fig. 8—Directional Characteristics of Several Wave-Antennas.

of these two types. In general, the directivity of dissimilar antennas may be increased with no loss in desired signal receptivity by combining them in arrays with little or no separation between the individual antennas. To obtain an increase in directivity by using several identical antennas in an array, however, without too great a sacrifice in desired signal receptivity, the array must cover a space comparable to and of the same order of magnitude as the wavelength of the signals for which it is designed.

It has been stated earlier in this paper that the fundamental form of wave-antenna consists of a single straight horizontal wire, terminated to ground at each end in its characteristic impedance. If the input circuit of a radio receiver be connected across the termination at the end of the antenna most distant from the desired transmitter, (the far end of the antenna) this simple form of wave-antenna can be used as a directional receiving system. If arrangements are made to bring the output from the initial end of the wave-antenna to the radio receiver as well as the output from the far end, the simple wave-antenna immediately becomes available for use as two identical antennas in an array. The ends of these two antennas from which the outputs are taken are separated by the length of the antenna and their axes are parallel but in the opposite sense. If before combining these two output currents, that from the initial end of the antenna is changed in phase and magnitude by the proper amount, it is possible to produce a null point of reception in any desired direction. The name "compensation" has been applied to the use of a single wave-antenna to form this array.[13] Since this null point is produced by balancing the back-end current from one antenna of the array (relative to its directional diagram) against the front-end current from the other antenna, the null point does not remain in the directional characteristic over a band of frequencies.

A directional diagram of a single antenna compensated to produce a null point at $\Theta = 161.4$ degrees (the bearing of the Rocky Point transmitter relative to the axis of the antenna) is shown in Fig. 10. This diagram was calculated, by the method outlined in Appendix 2, from the average of the measured constants of Houlton antennas *A*, *B*, and *D*. In this same figure, measured points are indicated, these points being the average of observations on these three antennas.

Beverage, Rice, and Kellog[13] have shown that there are important practical advantages to be gained by constructing the wave-antenna as a two-wire line, and using the metallic circuit acquired thereby as a transmission line to bring the output from one end of the antenna to the radio receiver. The circuits used to bring the output currents from the two ends of the wave-antenna to the radio receiver are shown in Fig. 11. In this case, the radio receiver is located at the initial end of the antenna, so that the predominant desired signal currents are transmitted over the metallic circuit of the wave-antenna to the radio receiver, while the compensation currents are taken directly from the initial end termination when this form of array is used.

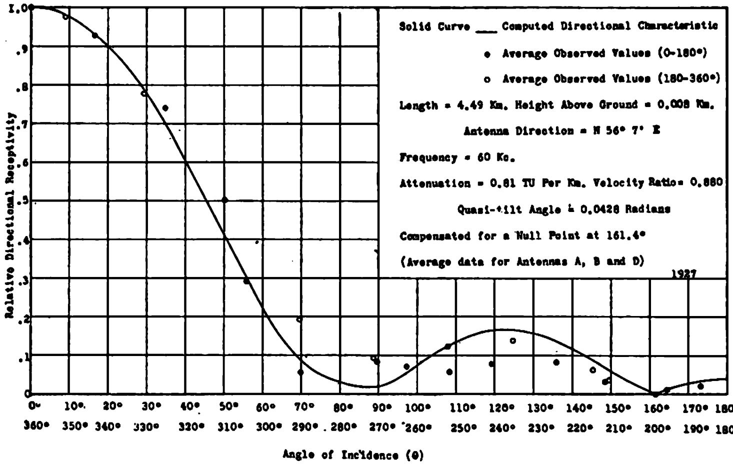

Fig. 10—Wave-Antenna Directional Characteristic. Relative Directional Receptivity of Compensated Average Houlton Antenna. (Short)

To obtain a greater reduction in the Noise Reception Factor (defined under "Directional Discrimination Against Static" earlier in this paper) than is given by compensation, two or more parallel wave-antennas are used in either a lateral array, a longitudinal array, or a combination of the two.

In the lateral array, the initial ends of the wave-antennas are spaced in the direction perpendicular to the axes of the antennas. Since it extends over space in the lateral direction, unless there be undue sacrifice in desired signal, the lateral array can only reduce the width of the directional diagram.

In a true longitudinal array, the antennas are coaxial, but their initial ends are separated by an appreciable fraction of a

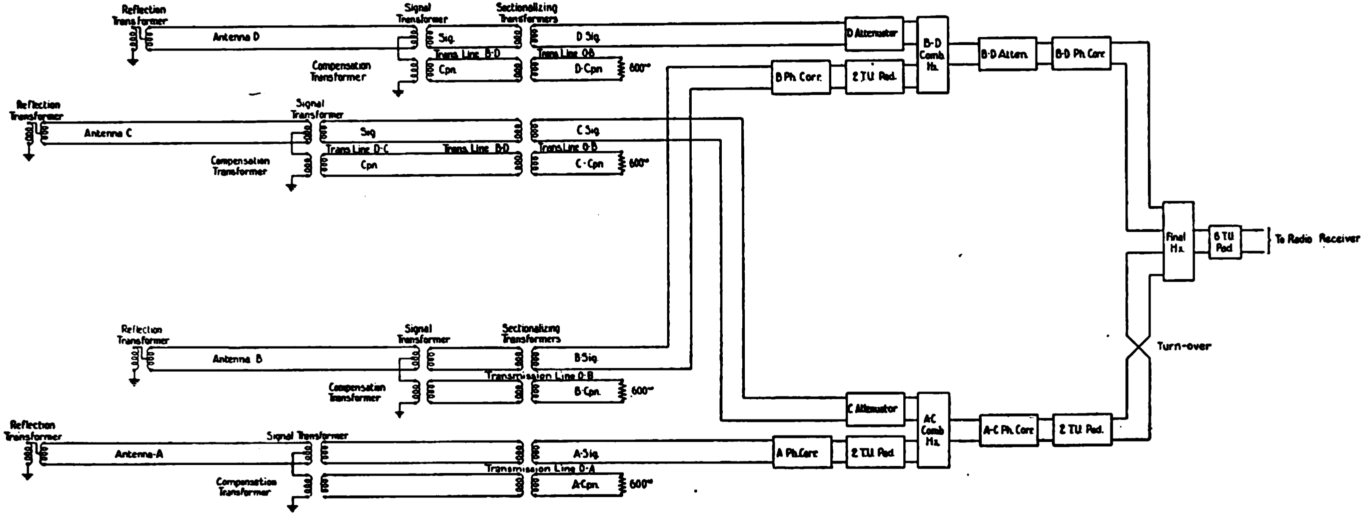

Fig. 11—Houlton Antenna System.

wavelength. If the wave-antennas forming this type of an array overlap one another, then the mutual impedance between them would greatly modify their individual characteristics. In practice, a small amount of lateral spacing between the units of a longitudinal array is necessary to make the mutual impedance negligible. When this type of array is properly designed, the reduction in directional receptivity due to the array is principally in the back-end direction.

The physical layout of the Houlton antenna system is shown in Fig. 12, and the circuits serving to connect the antennas to the radio receiver are shown in Fig. 11. The same letters are used for corresponding line sections in both of these figures. At the time that the directional characteristics of the Houlton antennas were measured, the antenna system comprised only three antennas, *A*, *B*, and *D*. Antenna *A* at that time extended from pole A-33 to pole A-117. Two arrays could then be used: antennas *B* and *D* forming a lateral array, and antennas *A* and *B* forming a modified longitudinal array. In normal operation using either of these two arrays, the transducers in the antenna output circuits were adjusted to combine equal amplitudes of the desired signals from the two antennas in phase with one another.

Using as the unit antenna for the arrays a directional diagram derived from the average constants of the antennas *A*, *B*, and *D*, the directional diagrams of these two arrays have been computed. Fig. 13 shows the computed directional characteristic of the lateral array and Fig. 14 the computed directional characteristic of the modified longitudinal array. On each of these figures, the measured points are shown.

The three antennas *A*, *B*, and *D* represented an uneconomical antenna system inasmuch as but two of the antennas could be used simultaneously in an array. To utilize fully these three antennas, at the same time increasing the discrimination against noise, the fourth antenna *C* has been constructed. To use these four antennas, they are arranged in pairs to form two lateral arrays, and the two lateral arrays arranged in a longitudinal array. The resultant total array quite evidently combines the narrowing of the directional diagram due to the lateral array and the reduction of the back end area of the directional diagram caused by the longitudinal array. A map of this array is shown in Fig. 12.

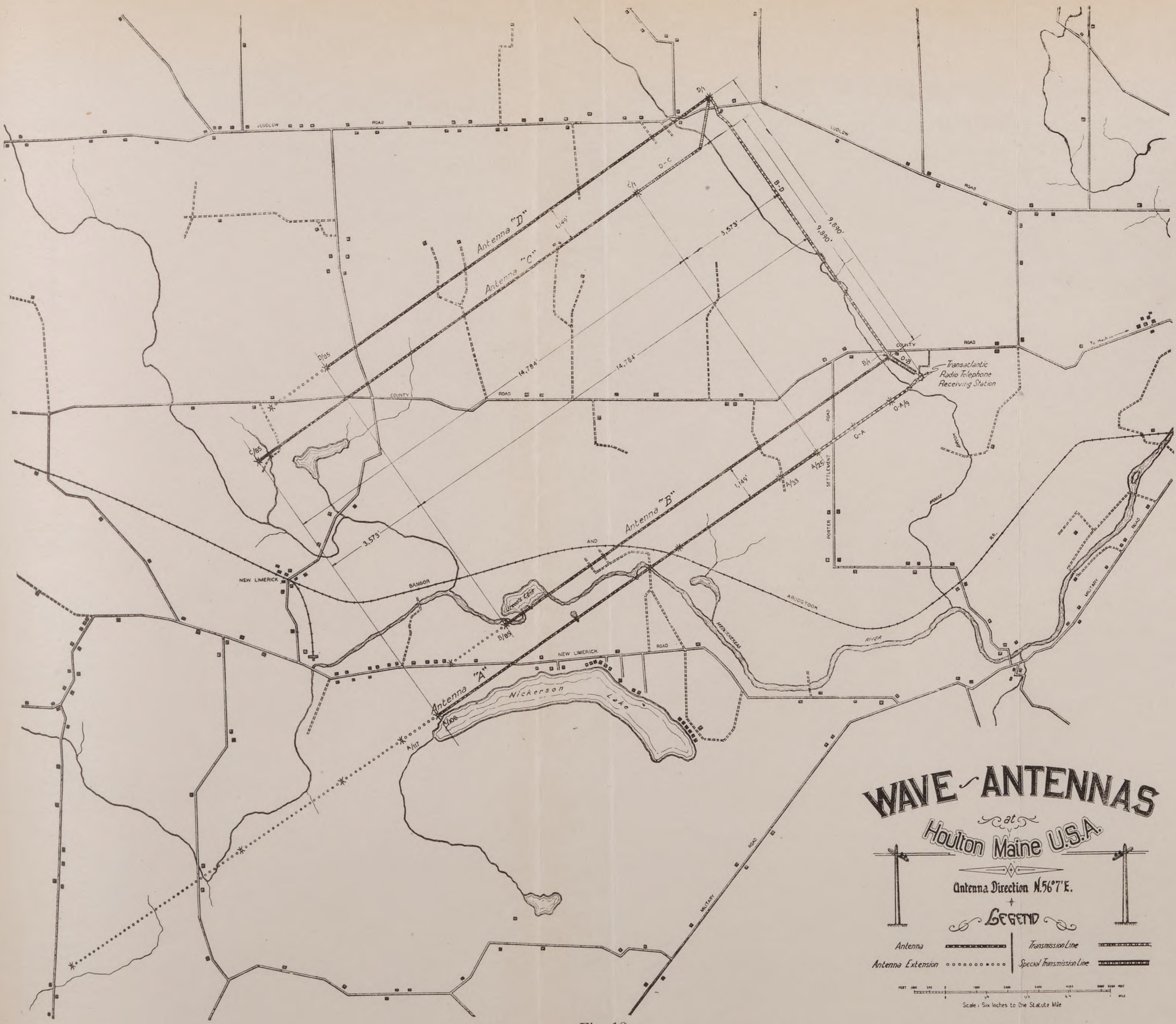LUDLOW
ROAD
LUDLOW
ROAD
D/1
D-C
C/1
B-D
9,890'
9,890'
Antenna "D"
Antenna "C"
1,149'
3,573'
14,784'
14,784'
D/85
C/85
COUNTY
ROAD
COUNTY
ROAD
B/1
O-B
Transatlantic
Radio Telephone
Receiving Station
O-A/9
O-A
A/25
A/33
1,149'
ROAD
SETTLEMENT
PORTER
Antenna "B"
BROOK
MOOSE
RR.
AND
NEW LIMERICK
BANGOR
AROOSTOOK
MEDUXNEKEAG
RIVER
Grecco's Lake
B/85
NEW LIMERICK
ROAD
Antenna "A"
A/108
A/117
Nickerson
Lake
MILITARY
ROAD
MILITARY
ROAD
WAVE ANTENNAS
at
Houlton Maine U.S.A.
Antenna Direction N.56°7'E.
LEGEND
Antenna
Antenna Extension
Transmission Line
Special Transmission Line
FEET 1000 500 0 1000 2000 3000 4000 5000 5280 FEET
0 1/4 1/2 3/4 1 MILE
Scale: Six Inches to One Statute Mile

Fig. 12

The circuits for combining four antennas of an array of the type described in the preceding paragraph are shown in Fig. 11. Antennas *A* and *C* form one lateral array; antennas *B* and *D* form the second. Since antennas *C* and *D* are further removed from the station than *A* and *B*, phase correctors are inserted in the circuits from *A* and *B* to compensate for the phase change in the transmission lines from *C* and *D*, so that the desired signals are combined in phase. The combination of the 2 TU fixed pads and the variable attenuators makes it possible to correct for the attenuation in the transmission lines to the more distant antennas. These several output currents are actually combined in hybrid coils, since this method of combination prevents the antennas from reacting one upon another through the combining system.

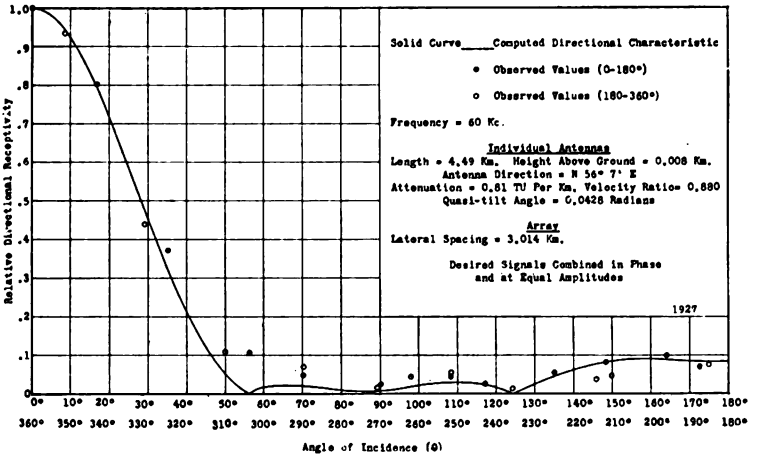

Fig. 13—Wave-Antenna Array Directional Characteristic. Relative Directional Receptivity of Lateral Array of Two Houlton Antennas. (Short)

After the antennas are combined in pairs to form two lateral arrays, the lateral arrays are combined in the longitudinal array.

The change of phase of space waves between one antenna and the next in an array is a linear function of frequency, and that on the metallic transmission lines practically so. By using phase correctors which have a phase change linear with frequency,[17] the outputs of the antennas in the array may then be combined to produce a null point or a reduction in receptivity, as a result of the array, which retains the same position in the directional diagram for every frequency within a finite band. The longitudinal array at Houlton is designed and combined to produce such an invariable null point in the direction 161.4 degrees relative to the axis of the wave-antenna array. At this angle of incidence, it is evident that the space waves arrive at the lateral array of antennas *A* and *C* before arriving at the lateral array of antennas *B* and *D*. To bring these undesired signals in phase, therefore, phase shift must be introduced into the output of the first of these arrays. Part of this phase shift is supplied by the metallic transmission lines and part by the phase correctors in the combining equipment. At this point, the desired signals remain in phase as the frequency is varied, so that a turn-over (reversal) inserted in the circuit to the lateral array of antennas *A* and *C* before the array is combined produces the null point which is invariable with variation of the frequency. Under these conditions, the phase of combination of the desired signals, incident at zero angle, varies as the frequency of the desired signals varies. To minimize the effect of this change in phase over the desired frequency band, the spacing of the antennas in the longitudinal array must be so chosen that the desired signals combine very nearly in phase at the middle of the frequency band. For that reason, the longitudinal spacing

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[17] O. J. Zobel, "Distortion Correction in Electrical Circuits with Constant Resistance Recurrent Networks," *Bell System Tech. Jour.*, 7, 438; July, 1928.

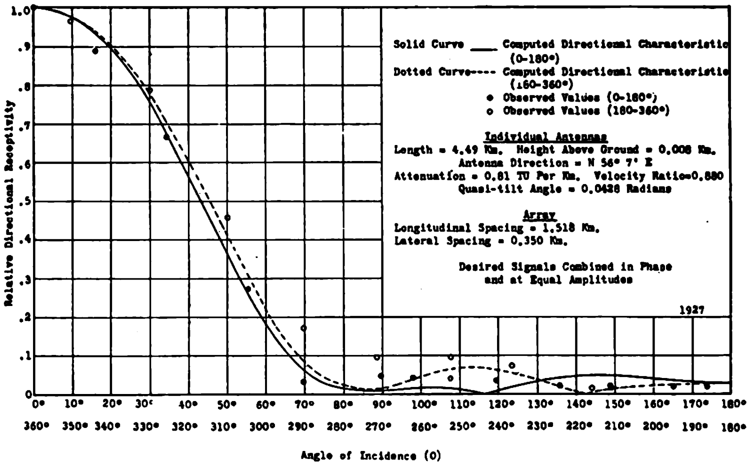

Fig. 14—Wave-Antenna Array Directional Characteristic. Relative Directional Receptivity of Modified Longitudinal Array of Two Houlton Antennas. (Short)

in the modified longitudinal array was decreased at the same time that the fourth antenna was constructed.

At the time that the extension of the antenna system was undertaken, the measured directional characteristics of the antennas *A*, *B*, and *D* were available, so that the unit directional diagram for use in determining this array characteristic was taken as the average measured characteristic of these three antennas. The calculated directional diagram of the complete Houlton antenna system is shown in Fig. 15. It should be noticed that the scale for the back-end dotted curve is ten times as great as that for the full-line curve for the major lobe.

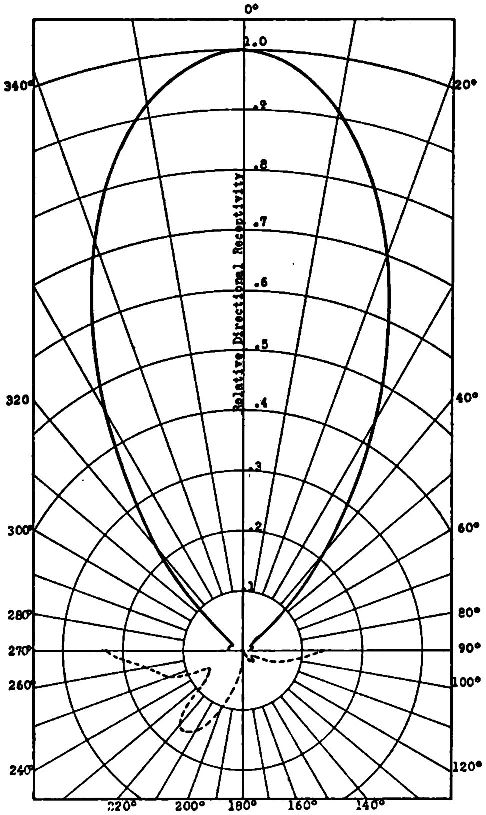

Fig. 15—Wave-Antenna Array Directional Characteristic. Calculated Relative Directional Receptivity of Array of Houlton Antennas *A*, *B*, *C*, and *D*. (From Average Measured Unit Antenna Characteristic.) Dotted Curve—Magnified ×10.

It is believed that the directional diagram, shown in Fig. 15, represents about the ultimate that can be done economically in a general reduction of back-end area and narrowing of the diagram by means of wave-antennas. Future extensions or redesign of the array at Houlton must be based on the reduction of the relative receptivity in distinct directions determined either by statistical study of the noise received by the antenna system or actual measurements of the direction of arrival of the noise which limits the operation of the transatlantic radio-telephone circuit.

THE RADIO RECEIVER

A description of the design and performance of the radio-telephone receiving set will constitute another paper. The radio receiving equipment employed in connection with the antenna systems was developed and constructed by Bell Telephone Laboratories.

The major transmission requirements upon which design must be based are as follows:

1. The limiting values of the signal field to be received.
2. The output power of the receiving antenna for a given signal strength.
3. Power output required from the radio receiver.
4. The type of telephone transmission to be received.
5. The frequency band to be received.
6. The nature and strength of interference from other radio stations and from noise. The selectivity required to reduce undesired modulation.
    - a. In amplifiers.
    - b. In demodulators.
7. Stability of frequency, gain and transmission-frequency characteristic.

1. Limiting Values of the Signal Field To Be Received. The range of daily averages of signal field at 60 kilocycles for all daylight-path hours is shown in Fig. 4. The fields, as previously published data indicate,[2,5,8] vary diurnally between much wider limits. At night the field frequently approaches, as a maximum, the value calculated on the basis of the inverse distance law. During sunrise and sunset dip periods the field frequently goes to a value less than one microvolt per meter with even 50 kilowatts radiated from a transmitter 5000 kilometers away. Suppose we take as being approximately correct values, field strengths of 0.4 microvolts per meter as the lower limit and 400 microvolts per meter as the upper limit. We then have determined that the receiving set should have a variation of gain of 60 TU.

2. The Output Power of the Receiving Antenna for a Given Signal Strength. From the observed constants of a Houlton wave-antenna and the assumed value of 0.4 microvolts per meter received at zero degree to the antenna direction as the lowest field, we calculate, using equations (125), (126), and (127) in Appendix 1, that the power supplied to the reflection transformer terminals is 3.716×10^{-6} microwatts. This power must suffer loss as a result of the transmission back to the receiving station over transmission lines and as a result of the necessity of providing flexibility in the operation of the apparatus used to combine the output of the antenna in question with the output of other antennas before it reaches the input terminals of the radio receiving set. (See Fig. 11.) This loss is such that the power at the input terminals of the radio receiving set from a single antenna and for the minimum signal field is very nearly equal to 3.7×10^{-7} microwatts. With the combining system actually used, the input to the radio receiver from all four antennas will be 12 TU above this value or 5.9×10^{-6} microwatts.

3. Power Output Required from the Radio Receiver. The value of output power required from the radio receiver is really governed by considering the whole radio circuit as a part of a long-distance telephone system. An overall loss of 10 TU has been found satisfactory for long toll circuits. If the telephone lines connecting the circuit terminals to the transmitting and receiving stations have an equivalent of 0 TU then we can place the 10 TU loss in the radio portion of the circuit. If we then supply on a single frequency within the voice-frequency band a power of 1 milliwatt to the input terminals of the radio transmitter, to get a 10 TU equivalent in the radio circuit we must obtain 0.1 milliwatt at the output of the radio receiver.

In the preceding section we determined that the minimum input would be 3.7×10^{-7} microwatts from a single antenna and hence the maximum gain required in the radio receiver to raise this power to the specified 100 microwatts output is 84 TU.

Within amplifiers using three-electrode vacuum tubes noise is generated in two ways: (a) by thermal agitation[18] in the conductor of the input circuit; and (b) by "Schottky Effect"[19] in the vacuum tubes themselves. Since the transatlantic radiotelephone circuit is so operated that the strength of the voice waves, or "electrical volume," is constant at the output of the radio receiver,[20] the maximum allowable noise at this point in the circuit is likewise constant. Good engineering practice specifies that the continuous "tube noise" should be more than 40 TU below the signal or less than 0.01 microwatt for the specified receiver output of 100 microwatts when using any gain up to the maximum of 84 TU. (With uniformly distributed noise over the voice-frequency band, this is equivalent to about 400 noise units.[21])

4. The Type of Telephone Transmission To Be Received. The "single-sideband, suppressed-carrier" type of telephone transmission, invented by John R. Carson,[22] has long been used in the Bell System in carrier systems on wire circuits.[23] Since the advantages of single sideband in radio transmission have been described by Hartley,[24] and in the radio transmitter by Heising,[25] we shall only briefly review the benefits arising from its use.

Transmission of two sidebands with the carrier suppressed represents an improvement over the "carrier and two-sideband" method ordinarily used in "broadcasting" since all of the transmitter power may be concentrated in the intelligence bearing frequencies. By transmitting only one sideband, further advantages are gained since the frequency space occupied is slightly more than halved for the same grade of circuit, the distortion at the output of the receiver is decreased, and practical simplifications may be made at the transmitting and receiving stations.[26] If the radio transmitter radiates equal power in each of the above-mentioned suppressed carrier transmission schemes and if the radio receiver accepts only the intelligence-bearing frequencies in each case, then the signal-to-noise ratio will be the same,[24] provided the resupplied carrier is in frequency synchronism in both systems and in addition in phase synchronism with the suppressed carrier in the two-sideband system.[26]

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[18] J. B. Johnson, "Thermal Agitation of Electricity in Conductors," *Phys. Rev.*, 32, 97; July, 1928.

Harry Nyquist, "Thermal Agitation of Electric Charge in Conductors," *Phys. Rev.*, 32, 110; July, 1928.

J. B. Johnson, "Thermal Agitation of Electricity in Conductors," *Nature*, 119, p. 50; Jan. 8, 1927.

[19] Walter Schottky, "Atomare Schwankungsvorgänge an Glühkathodenoberflächen," *Physik. Zeitschr.*, 27, 701; Nov. 1, 1926.

T. C. Fry, "The Theory of the Schroteffekt," *Jour. Frank. Inst.*, 199, 203; Feb., 1925.

J. B. Johnson, "The Schottky Effect in Low Frequency Circuits," *Phys. Rev.*, 26, 71; July, 1925.

[20] S. B. Wright and H. C. Silent, "The New York-London Telephone Circuit," *Bell System Tech. Jour.*, 6, 736; October, 1927.

[21] The noise unit is an arbitrary unit used in the Bell System for comparison of any noise with a certain arbitrary source of noise known as a noise standard. The output of the noise standard may be attenuated to produce the same interfering effect on speech as the noise being measured. See W. H. Harden, "Practices in Telephone Transmission Maintenance Work," *Bell System Tech. Jour.*, 4, 26; January, 1925, for details of making such comparisons.

[22] U. S. Patents Nos. 1,343,306 (1920); 1,343,307 (1920); 1,449,382 (1923), to J. R. Carson.

[23] E. H. Colpitts and O. B. Blackwell, "Carrier Current Telephony and Telegraphy," *Trans.* A.I.E.E., 40, 205; 1921.

[24] R. V. L. Hartley, "Relation of Carrier and Sideband in Radio Transmission," PROC. I. R. E., 11, 34; Feb., 1923.

[25] R. A. Heising, "Production of Single Sideband for Transatlantic Radio Telephony," PROC. I. R. E., 13, 291; June, 1925.

When receiving single-sideband transmission the carrier suppressed at the transmitting station is resupplied in the radio receiver. Since this carrier will demodulate both sidebands with equal efficiency, the opposite sideband must be eliminated before demodulation to prevent the noise in this sideband from appearing in the voice-frequency output. If the noise power in either sideband is p, then the noise power without the opposite sideband suppression is $2p$ and if we reduce the noise power from the opposite sideband to $0.1p$ the total received noise will be reduced

$$10 \log_{10} \frac{2p}{p+0.1p} = 2.59\text{TU}$$

The maximum possible reduction in noise is 3.01 TU,[27] so that for engineering purposes a 10 TU suppression of the noise in the opposite sideband may be considered adequate. For other reasons to be brought out later in this paper the opposite sideband loss must be greatly in excess of this value.

Provided the resupplied carrier used to demodulate the single sideband suppressed carrier signals is large relative to the signal magnitude[24] at that point in the circuit where we choose to supply it, the only other requirement is that its frequency be correct. Since a displacement of the resupplied carrier 50 cycles above or 20 cycles below the zero of the equivalent voice-frequency band is sufficient to give an appreciable decrease in speech intelligibility, its frequency should be maintained within the smaller of these two limits or within plus or minus 20 cycles of the correct value. It is interesting to note that an absolute variation of only one-tenth of this amount can be observed on music and that speech naturalness is similarly affected.

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[26] J. R. Carson, "Signal-to-Static-Interference Ratio in Radio Telephony," Proc. I. R. E., **11**, 271; June, 1923.

[27] J. R. Carson, "Selective Circuits and Static Interference," *Bell System Tech. Jour.*, **4**, 265; April, 1925.

5. Frequency and Frequency Band To Be Received. To utilize the power available at the transmitter most effectively, it is essential to transmit only those frequencies contributing most to received intelligibility. The energy of speech lies largely below 500 cycles while the frequencies most important for intelligibility lie between 400 and 2600 cycles.[28] By limiting the band transmitted to speech frequencies above 400 cycles some saving is obtained in the transmitter power required. The range of frequencies transmitted may then extend from 58.9 to 61.1 kilocycles with the suppressed carrier at 58.5 kilocycles, and the radio receiver must be designed to accept this band of frequencies. The transmission-frequency characteristic of the overall radio receiving set should not vary more than ±2TU within the band specified above to give a good telephone communication circuit.

6. Selectivity Requirements. The selectivity required in the receiving set is such that when the desired signal is at the assumed minimum value no deleterious effects will be caused by undesired signals.

In Fig. 16 there are shown measured daylight field strengths of various existing radio-telegraph stations as observed at Riverhead, New York, and at Cupar, Scotland. Since these measurements could not be indefinitely extended in frequency nor could they take into account all stations which might exist in this range, they may be considered only as a guide in obtaining a curve of the maximum telegraph interference to be expected. These data, in Fig. 16, have been expressed as ratios (in TU) to the minimum desired signal to be received. It is important to note that the directional selectivity of the receiving antenna system used materially decreases the relative magnitude of many of these interfering signals, particularly at the higher frequencies. The American receiving station is now located in Houlton, Maine, instead of Riverhead, New York, and this increase in distance from many of the American high-power transmitting stations decreases somewhat their field strengths. In view of these factors the "Assumed Maximum Interference," although it is not greater than any observed station field strength, is, in fact, greater than the interfering signals when the outputs from the actual receiving antennas are used instead of field-strength observations.

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[28] W. H. Martin and Harvey Fletcher, "High Quality Transmission and Reproduction of Speech and Music," *Trans.* A.I.E.E., 43, 384; 1924.

6a. Selectivity Requirements Imposed by the Use of Amplifiers. To operate a vacuum tube as an amplifier with negligible distortion the peak voltage applied to its grid must be less than a limiting value so that the tube always operates over the practically linear portion of its characteristic. If no discrimination were provided against unwanted signals, we would be placed in the peculiar situation of having to supply ample tube capacity in the radio receiver to care for the combined load produced by perhaps 100 telegraph stations each of which, on the average, may have a received signal strength 1000 times the assumed minimum signal. An easy way to decrease the load produced by interference is to insert a filter at the input of the receiving set, which will reduce the required capacity of the first tube. Additional selectivity following the first tube still further reduces the load of undesired signals on the following tubes as more of the capacity of those tubes is used for the desired signals.

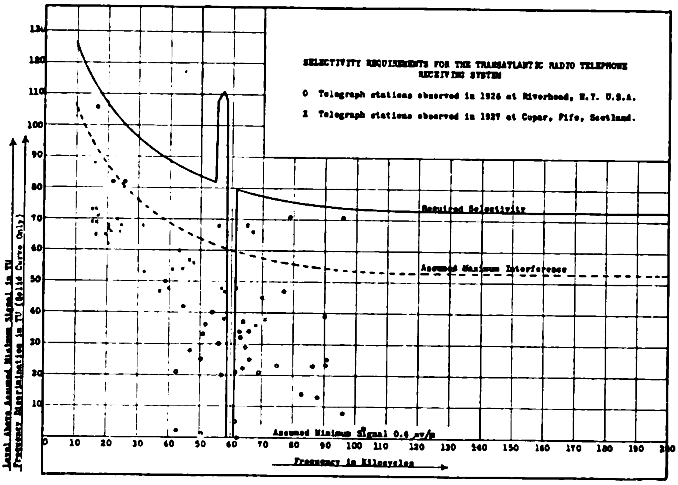

Fig. 16.

Now for design purposes let us assume that the load capacity of each tube is at least 6 TU greater than the capacity required in the tube for the performance of its functions on the desired signal. The undesired signals may then be allowed to produce on the tube grid a voltage equal to that of the desired signal.

Since each of the undesired signals shown in Fig. 16 are about 60 TU stronger than the minimum desired signal, they must be reduced by that amount to make them each no greater than the desired signal.

It is shown in Fig. 21 of Appendix 4 that, as a result of unit random input voltages from 100 operating radio-telegraph stations, a peak voltage will be produced equal to or greater than 10 such units during less than 0.1 per cent of the time. If the undesired telegraph station signals were all of the same magnitude as the desired signal then the voltage which they would produce would be 20 TU above the voltage of the desired signal.

From purely load considerations then, the total required suppression of every interference-bearing frequency outside of the desired signal receiving band will be

$$60+20=80 \text{ TU}$$

6b. Selectivity Requirements Imposed by the Use of Demodulators. In all demodulators, the range of desired output frequencies should not be included in the input frequency band because the input frequencies amplified appear in the output of the demodulator as the first order modulation product. The output band of the demodulator should be at a lower frequency than the input band in order to reduce the number of undesired modulation products in the output and in order to obtain the benefit of greater selectivity from circuits operating at lower frequencies. Of course, if all the required selectivity can be conveniently put before the first demodulator, there is no valid reason why multiple-demodulation should be used.

In all demodulators except the final demodulator of a radio receiver, the band of frequencies allowed to pass into the demodulator should not be greater in width than the absolute value of the lowest desired frequency in the demodulator output. This requirement is apparent when we consider second order modulation products of interference within the band accepted by the demodulator. Suppose we assume the use of double-demodulation and choose 30 kilocycles as the lowest desired frequency in the output of the first demodulator. Then if the band impressed upon the demodulator be more than 30 kilocycles in width, two interfering signals within the band might together give a difference frequency of 30 kilocycles producing load in subsequent stages and possibly tone or noise in the output circuit.

Second order modulation between two signals, one lying within the band accepted by the demodulator and one outside it, may also give rise to interference, due to the difference frequency falling in the output band of the demodulator. Assume that one interfering signal lies within the band accepted by the demodulator and is +60 TU referred to the minimum desired signal at the grid of the first tube. An equal signal at a frequency outside the band and subject to the selectivity provided for meeting the load requirement will be −20 TU referred to minimum desired signal at the same point. Since the second order output from a demodulator is approximately proportional to the product of the grid voltages producing it,[29] we may write (in TU):

Relative desired signal = (0) + (Beating oscillator voltage)
Relative interference = (+60) + (−20) = +40

Tests have shown that an interrupted tone, similar to telegraph interference, which is heard at a frequency of 1100 cycles in a telephone receiver is about the most serious frequency of interference to received speech on a telephone circuit, and that frequencies above and below 1100 cycles are of somewhat less importance. Signals at the equivalent 1100-cycle frequency produce a type of interference which good engineering practice requires should be reduced at least 50 TU below the desired signal. (This amount of interference is equal to about 500 noise units at the −10 TU transmission level.[21])

To satisfy this requirement, the relative desired signal should be 50 TU greater than the relative interference at the output of the demodulator. Assigning a minimum magnitude to the beating oscillator voltage of 90 TU above the minimum desired signal voltage on the grid of the demodulator reduces this type of interference sufficiently. (Using a balanced demodulator arrangement, this value might be reduced some 20 TU.)

Since any two signals at frequencies entirely outside the band accepted by the demodulator are suppressed some 80 TU, we need not consider their second-order modulation products.

If we use double demodulation in a radio receiver as is assumed in the first part of this section, then we must consider other products of modulation with the beating oscillator for frequencies distant from the accepted band. Space will not permit us to more than mention these, but since the frequencies to be suppressed are distant from the frequencies to be received their suppression is relatively simple.

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[29] J. R. Carson, "A Theoretical Study of the Three-Element Vacuum Tube," PROC. I. R. E., 7, 187; April, 1919.

In the final demodulator of a radio receiver we must tolerate a certain amount of distortion due to the intermodulation of input frequencies. By limiting the band width into the final demodulator to the same width as the desired output band, the distortion due to intermodulation with interference lying outside the desired band is eliminated. By supplying a large amount of carrier to the final demodulator and by using a balanced demodulator the amount of noise and distortion due to intermodulation of frequencies lying inside the desired band is reduced.

If the desired signal band extends from 58.9 to 61.1 kilocycles and is an upper sideband corresponding to voice frequencies from 400 to 2600 cycles then, in effect, we must supply a carrier at 58.5 kilocycles to produce the proper voice frequencies in the output circuit. This carrier frequency will also demodulate the frequencies below it in such a way as to produce audible signals and for this reason ample protection must be supplied against the opposite sideband if stations are likely to exist in that range. Calculations show that this is the case; for if 100 stations are distributed at random over the 190-kilocycle range between 10 and 200 kilocycles, then the probability that at least one station lies between 55.5 and 58.5 kilocycles is 0.796. If the assumed maximum interference at the equivalent 1100 cycles in the opposite sideband, as indicated by the dashed line in Fig. 16, is 61 TU above the minimum signal and we wish to have it 50 TU below, as previously stated, then we require a selectivity of 61 TU plus 50 TU or 111 TU for this frequency. For other tone-producing frequencies of the opposite sideband similar selectivity requirements have been set up and the resultant for frequencies from 55.5 to 58.5 kilocycles is shown by the solid curve in Fig. 16.

7. *Stability.* As mentioned in Section 4 above, the carrier for a single sideband receiver must be resupplied at the correct frequency. All of the oscillators in the radio link must have sufficient frequency stability to maintain the voice frequencies at the receiver output correct within 20 cycles per second over long periods of time. Suppose we allow 10 cycles per second variation in frequency to exist at the transmitter and an equal amount at the receiver, then the variation in frequency at the receiver

must never exceed 0.017 per cent if the resupplied carrier is at 58.5 kilocycles. Certain advantages in stability of the resupplied carrier can be obtained by the use of double demodulation in the receiving set and these will be discussed in another paper.

Variations in the efficiency of the transatlantic radio transmission path for long wavelengths occur with time of day and season, but during any individual all-daylight transmission period the transmission efficiency of the path is fairly constant. If the gain of the receiver is constant, then, during this important period of the day, the minimum of circuit adjustments will be required. It is hence desirable that the gain of the entire receiving set be made to hold constant within ±2 TU for all variations of temperature and of voltage of battery supply, within the operating limits.

It is almost self-evident that the transmission-frequency characteristic through the radio receiver should not vary with temperature and time. Changes of this nature should not exceed 0.5 TU within the transmission band nor 5 TU outside of the transmission band. Design of stable filters and vacuum-tube circuits are essential to produce this result.

The authors have endeavored, in the limited space of the preceding pages, to show what radio transmission considerations must be taken into account in properly designing a receiving system for a commercial radio-telephone circuit. A rather detailed discussion has been necessary to present an accurate picture of the various factors entering into the production of the very essential and highly directional long-wave receiving antenna system employed.

The cooperation of the engineers of the Wireless Section of the British General Post Office, particularly Col. A. G. Lee and Mr. I. J. Cohen, in the measurements made on wave-antennas in England and Scotland, is greatly appreciated and we take this occasion to thank them for having made possible the obtaining of these data. All of our early work in connection with wave-antennas and our initial field trials of lateral and longitudinal arrays of wave-antennas were carried out using wave-antennas located at Belfast, Maine, and Riverhead, New York. These antennas were made available through the courtesy of the Radio Corporation of America, and the authors wish to express to Mr. H. H. Beverage of that organization their appreciation for his interest and assistance during the tests.

Appendix 1

THE WAVE-ANTENNA

Fundamentally, the wave-antenna consists of a straight horizontal wire, terminated to ground at each end in its characteristic impedance.[13] The determination of the receptivity characteristics of the wave-antenna consists in determining the current flowing in the terminal impedances of the antenna resulting from a field impressed along the antenna.[30]

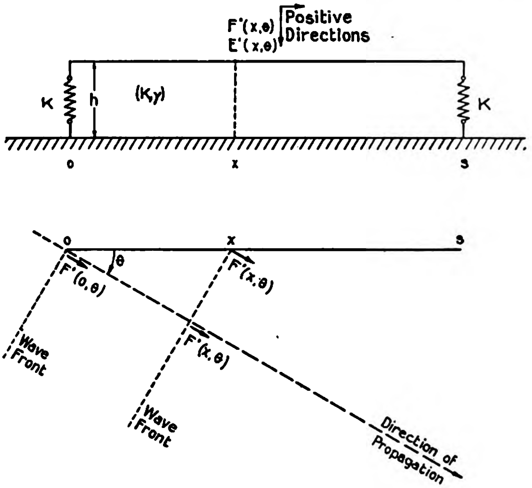

Fig. 17.

The wave-antenna is shown in Fig. 17, consisting of a line of length s extending from $x=0$ to $x=s$. In the nomenclature of the following discussion, letters with no primes refer to the antenna, letters with a single prime (′) to the impressed field, and letters with a double prime (″) to the resultant field. The wave-antenna is in an impressed electromagnetic field which is defined by the quantities ϕ', V', f_w', and f_g' where

ϕ' = impressed magnetic flux between the lower surface of the wire and the surface of the ground (per unit length).

V' = impressed electric force between the wire and the ground.

f_w' = the impressed electric force *along* the lower surface of the wire.

f_g' = the impressed electric force *along* the surface of the ground.

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[30] J. R. Carson and R. S. Hoyt, "Propagation of Periodic Currents over a System of Parallel Wires," *Bell System Tech. Jour.*, **6**, 495; July 1927.

The total field about the antenna is the sum of this impressed field and a secondary field due to the currents and charges produced in the circuit by the impressed field, so that

$$\begin{aligned} \phi'' &= \phi' + \phi \qquad & f_w'' &= f_w' + f_w \\ V'' &= V' + V \qquad & f_g'' &= f_g' + f_g \end{aligned} \tag{101}$$

where ϕ, V, f_w, and f_g are the components of the secondary field set up by the currents and charges in the system and ϕ'', V'', f_w'' and f_g'' represent the resultant field about the system.

As a result of the impressed field, a current I flows in the wire, and a corresponding superposed current distribution is induced in the ground. If the internal impedance of the wire be z_w and that of the ground be z_g, the resultant longitudinal electric force along the wire may be written

$$f_w'' = Iz_w = f_w' + f_w \tag{102}$$

and similarly the resultant longitudinal electric force along the ground is

$$f_g'' = (-Iz_g + f_g') = f_g' + f_g \tag{103}$$

The second curl law applied to the periphery of the rectangle formed by the vertical at x, the wire, the vertical at $(x + \Delta x)$, and the ground yields

$$zI - f_g' + \frac{dV''}{dx} = -\frac{d\phi''}{dt} \tag{104}$$

where z is the total series impedance of the wire and the ground circuit and is

$$z = z_g + z_w. \tag{105}$$

A summation of the voltages around the above defined rectangle yields

$$f_w' - f_g' + \frac{dV'}{dx} = -\frac{d\phi'}{dt} \tag{106}$$

Subtracting (106) from (104) we get

$$zI - f_w' + \frac{dV}{dx} = -\frac{d\phi}{dt} \tag{107}$$

If we write Q as the charge, C as the capacity to ground, and L as the external inductance, each per unit length of the wire, equation (107) becomes

$$zI+L\frac{dI}{dt}+\frac{1}{C}\ \frac{dQ}{dx}=f_w' \tag{108}$$

but the line current is decreased by the amount of the charging current and the leakage current

$$-\frac{dI}{dx}=\frac{dQ}{dt}+I_Y \tag{109}$$

where I_Y is the leakage current per unit length of the wire. If the admittance of the leak to ground be designated as Y, the leakage current is

$$I_Y=YV''=Y(V'+V) \tag{110}$$

Since we are interested only in the steady state, the operator d/dt may be replaced by $j\omega$. Substituting the expression (110) for I_Y into (109) and differentiating with respect to x yields

$$-\frac{d^2I}{dx^2}=\frac{dQ}{dx}j\omega+Y\left(\frac{dV'}{dx}\right)+\frac{Y}{C}\ \frac{dQ}{dx} \tag{111}$$

By means of (111) we may eliminate Q from (108)

$$(z+jL\omega)I-\frac{1}{Y+jC\omega}\ \frac{d^2I}{dx^2}=f_w'+\frac{Y}{Y+jC\omega}\ \frac{dV'}{dx} \tag{112}$$

and if

$$K=\sqrt{\frac{z+jL\omega}{Y+jC\omega}} \tag{113}$$

$$\gamma=\sqrt{(z+jL\omega)(Y+jC\omega)} \tag{114}$$

where K is the characteristic impedance and γ the propagation constant of the antenna circuit, equation (112) may be written

$$\frac{K}{\gamma}\left(\gamma^2-\frac{d^2}{dx^2}\right)I=f_w'+\frac{YK}{\gamma}\ \frac{dV'}{dx} \tag{115}$$

When the boundary conditions are applied, equation (115) defines the value of the current I in the wave-antenna in terms of the impressed electromagnetic field specified by V' and f_w'. By equation (101) the resultant voltages at the ends of the antenna are:

$$V''(0)=V'(0)+V(0) \tag{116}$$

$$V''(s)=V'(s)+V(s) \tag{117}$$

To this point, the solution of the wave-antenna problem has been in a rigidly analytic form. While it is possible to determine completely the received current by following through this method of solution, the problem can be greatly simplified and a physical picture of the problem gained by a synthetic process.

The synthetic method of attack consists of replacing the impressed field by a set of electromotive forces identically equivalent to the impressed field in the sense that it produces the same currents and charges.[30]

The proposed set of electromotive forces is as follows:

A. A distributed longitudinal electromotive force f_w' per unit length in the wire, i.e., an electromotive force $f_w'dx$ in each element of length dx.

B. A distributed vertical electromotive force, V', in the superposed shunt admittance Y between the wire and ground, i.e., an electromotive force V' in each elemental admittance path Ydx.

C. In each end of the wire, $x=0$ and $x=s$, localized series electromotive forces, equal respectively to minus and plus the impressed voltages at those points; i.e., equal to $-V'(0)$ and $+V'(s)$ respectively.

The electromotive force of A is suggested by (107), that of B by (109) and (110), that of C by the terminal conditions expressed in (116) and (117). In the case of a wave-antenna constructed to maintain high insulation resistance, the conductance portion of the superposed admittance Y can be made negligibly small. Under this condition, the susceptance part of this admittance can be combined with the linear capacitance of the wire to alter the propagation constants (K and γ) of the antenna and the voltages induced in the superposed shunt admittances neglected.

By reference to Fig. 17, the impressed field may be identically defined at each point along the antenna.

The longitudinal electromotive force in each element of the wire is

$$\begin{aligned} f_w'dx &= F'(x,\theta)\cos\theta\,dx \\ f_w'dx &= F'(0)\epsilon^{-\gamma' x\cos\theta}\cos\theta\,dx \end{aligned} \tag{118}$$

The impressed voltage at the point x along the antenna is

$$\begin{aligned} V'(x) &= h \cdot E'(x, \theta) \\ V'(x) &= h \cdot E'(0) \epsilon^{-\gamma' x \cos\theta} \end{aligned} \tag{119}$$

In (118) and (119) $F'(0)$ and $E'(0)$ represent the horizontal and vertical components respectively of the impressed electric field at the end of the antenna $x=0$, and h represents the height of the antenna above ground. For the purpose of this discussion, it will be assumed that F' and E' are not dependent upon θ. The current produced at the receiving end s by the horizontal component of the impressed field is given by

$$I_{F'\theta} = \int_0^s \frac{F'(0)\epsilon^{-\gamma' x \cos\theta} \cos\theta \, dx}{2K} \epsilon^{-\gamma(s-x)} \tag{120}$$

From which

$$I_{F'\theta} = \frac{sF'(0)\cos\theta}{2K} \frac{\epsilon^{(\gamma-\gamma'\cos\theta)s} - 1}{(\gamma-\gamma' \cos\theta)s} \epsilon^{-\gamma s} \tag{121}$$

The current produced at the receiving end s by the vertical component of the impressed field is evaluated as follows:

$$I_{E'\theta} = \frac{V'(s)}{2K} - \frac{V'(0)}{2K} \epsilon^{-\gamma s} \tag{122}$$

and by combination of (119) and (122)

$$I_{E'\theta} = \frac{hE'(0)}{2K} \left[\epsilon^{(\gamma-\gamma'\cos\theta)s} - 1 \right] \epsilon^{-\gamma s} \tag{123}$$

Zenneck's theory of wave propagation[31] has been developed by Breizig[32] to show that the horizontal and vertical components of the impressed field are related by the expression

$$-\frac{F'}{E'} = \epsilon^{j\delta} \tan T. \tag{124}$$

The total current produced at the receiving end s by the impressed field is

$$I_\theta = I_{F'\theta} + I_{E'\theta} \tag{125}$$

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[31] J. Zenneck, "Ueber die Fortpflanzung ebener electromagnetischer Wellen längs einer ebenen Leiterfläche und ihre Beziehung zur drahtlosen Telegraphie," *Ann. der Phys.*, 23, 846; June, 1907.

[32] Franz Breizig, "Theoretische Telegraphie," Braunschweig, 1924. 2nd ed., pp. 482–487.

and by application of (124) the constituents of the total current are

$$I_{F'\theta}=\frac{S\lambda'F'}{2K}\cos\theta\frac{1-\epsilon^{-[\alpha S\lambda'+j2\pi S(m-\cos\theta)]}}{\alpha S\lambda'+j2\pi S(m-\cos\theta)}\epsilon^{-j2\pi S\cos\theta} \tag{126}$$

$$I_{E'\theta}=-\frac{S\lambda'F'}{2K}\frac{h}{S\lambda'}\frac{1}{\epsilon^{j\delta}\tan T}$$

$$\left(1-\epsilon^{-[\alpha S\lambda'+j2\pi S(m-\cos\theta)]}\right)\epsilon^{-j2\pi S\cos\theta} \tag{127}$$

In (125), (126), and (127), the symbols have the following meanings

SYMBOL	DEFINITION	UNIT
I_θ	The total current produced at the receiving end of the antenna *s* by an impressed field propagated at an angle θ from the axis of the antenna.	amperes
$I_{F'\theta}$	The portion of I_θ produced by the horizontal component of the impressed field.	amperes
$I_{E'\theta}$	The portion of I_θ produced by the vertical component of the impressed field.	amperes
F'	The horizontal component of the impressed field. (Positive direction in the direction of propagation along the ground.)	volts per kilometer
E'	The vertical component of the impressed field. (Positive direction downward.)	volts per kilometer
δ	Phase angle between the horizontal and vertical components of the impressed electric field.	radians
T	"Quasi-tilt angle" of the impressed electric field.	radians
K	The characteristic impedance of the wave-antenna.	ohms
γ	The propagation constant of the wave-antenna.	
α	The real part of the propagation constant of the wave-antenna or the attenuation constant.	napiers per kilometer
β	The imaginary part of the propagation constant of the wave-antenna or the phase constant.	radians per kilometer
γ'	The propagation constant of the space waves.	
α'	The real part of the propagation constant of the space waves (assumed equal to zero).	napiers per kilometer

β'	The imaginary part of the propagation constant of the space waves.	radians per kilometer
s	The length of the wave-antenna.	kilometers
h	The height of the wave-antenna above ground.	kilometers
$S=s/\lambda'$	The length of the wave-antenna.	space wave-lengths
$\lambda'=2\pi/\beta'$	The wavelength of the space waves.	kilometers
$V=\dfrac{2\pi f}{\beta}$	Apparent velocity of propagation of waves along the wave-antenna.	kilometers per second
V'	The velocity of propagation of the space waves $(=3\times10^5$ km per second)	kilometers per second
V/V'	Velocity ratio.	numeric
$m\equiv V'/V=\beta/\beta'$	Reciprocal of the velocity ratio.	numeric
$j=\sqrt{-1}$		
θ	The angle between the axis of the wave-antenna and the direction of propagation of space waves measured in a clock-wise direction.	
$\text{R.D.R.}=\dfrac{I_\theta}{I_0}$	Relative directional receptivity.	numeric

Appendix 2

ANTENNA ARRAYS

The directional discrimination yielded by a single antenna can be increased by utilizing several such antennas in an array.[16] In

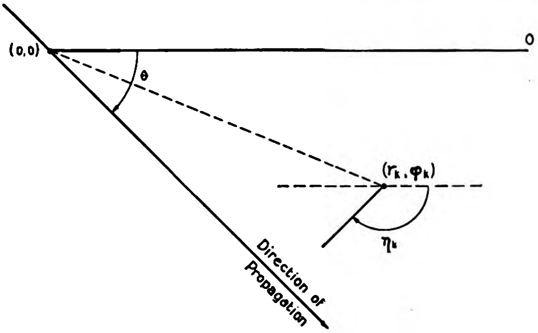

Fig. 18.

Fig. 18, a general array of n antennas is indicated, of which only the first and the k'th are portrayed.

Each antenna in the array is completely specified by the coordinates of the initial end of the antenna, the angle between the zero axis of the coordinate system and the axis of the antenna, and the current delivered at the receiving end of the antenna for a given electric field impressed on the antenna at each angle of incidence with the antenna. Literally, the first and the *k*'th antennas are specified as follows:

	First Antenna	*k'th Antenna*
Coordinates of initial end of antenna	$(0,0)$	(r_k,ϕ_k)
Direction of antenna	0	η_k
Current delivered by antenna for a constant electric field propagated in the direction θ	I_θ	$I_{\theta k}$

For the purpose of this discussion, it is sufficiently accurate to assume that the propagation of space waves over the area covered by the array only involves phase retardation, i.e.,

$$\gamma' = j\beta' \tag{201}$$

The output of the *k*'th antenna is transmitted through a linear transducer having a transfer constant P_k to a common point where it is combined with the outputs of the other antennas of the array. The current from the *k*'th antenna at the point of combination is therefore

$$J_{k\theta} = I_{\theta k}\epsilon^{-j\beta'[r_k/V']\cos(\theta-\phi_k)}\epsilon^{-P_k} \tag{202}$$

where

$$\theta_k = \theta - \eta_k \tag{203}$$

and

$$\beta' = \frac{2\pi V'}{\lambda'} \tag{204}$$

The total current received from the *n* antennas of the array is equal to the sum of the currents received from the individual antennas, or

$$J_\theta = \sum_{k=1}^{k=n} I_{\theta k}\epsilon^{-j[2\pi r_k/\lambda']\cos(\theta-\phi_k)}\epsilon^{-P_k} \tag{205}$$

Equation (205) gives the total current received from *any* array of antennas for any direction of wave propagation in a horizontal plane. This general expression is not adapted to ready determination of directional characteristics of antenna systems, but it may be simplified by placing the following restrictions on the individual antennas forming the array and their space relations in the array:

(1) The antennas are all alike. This restriction may be defined by the expression:

$$I_{\theta k} = I_{\theta(k+1)}$$

(2) The axes of the antennas are parallel, as defined by the expression

$$\eta_k = 0 \text{ or } \pi$$

(3) The initial ends of the antennas are equally spaced along straight lines in each sub-group and the sub-groups are equally spaced along straight lines. All of the sub-groups are identical.

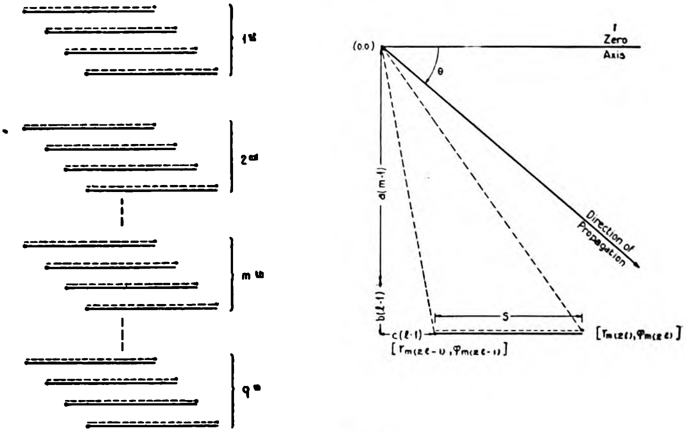

Fig. 19.

The general antenna array conforming to these restrictions is shown in Fig. 19. In this figure, there are q groups of antennas equally spaced by the distance a along a line 90 deg. from the zero axis. In each of these q groups of antennas, there are p antennas, divided into two series, those for which $\eta = 0$ being numbered 1, 3, $\cdots$, $(2l-1)$, $\cdots$ $(p-1)$ and those for which $\eta = \pi$ being numbered 2, 4, $\cdots$, $2l$, $\cdots$, p, the initial ends of the second series being removed by a distance s from the initial ends of the first series, along the axes of the antennas of the first series.

Equation (205) applied to this general array gives for the total current received from the array

$$J_\theta = I_\theta \sum_{m=1}^{m=q} \sum_{2l=2}^{2l=p} \epsilon^{-j\frac{2\pi r_{m(2l-1)}}{\lambda'}\cos(\theta-\phi_{m(2l-1)})} \epsilon^{-P_{m(2l-1)}}$$

$$+ I_{\theta-\pi} \sum_{m=1}^{m=q} \sum_{2l=2}^{2l=p} \epsilon^{-j\frac{2\pi r_{m(2l)}}{\lambda'}\cos(\theta-\phi_{m(2l)})} \epsilon^{-P_{m(2l)}} \tag{206}$$

where

$$r_{m(2l-1)} = \sqrt{[a(m-1)+b(l-1)]^2+[c(l-1)]^2} \tag{207}$$

$$r_{m(2l)} = \sqrt{[a(m-1)+b(l-1)]^2+[c(l-1)+s]^2} \tag{208}$$

$$\phi_{m(2l-1)} = \tan^{-1}\left[\frac{a(m-1)+b(l-1)}{c(l-1)}\right] \tag{209}$$

$$\phi_{m(2l)} = \tan^{-1}\left[\frac{a(m-1)+b(l-1)}{c(l-1)+s}\right] \tag{210}$$

In a double summation, the result is independent of the order in which the summations are taken. If then we write

$$u_\theta = \sum_{2l=2_{(m=1)}}^{2l=p} \epsilon^{-j\frac{2\pi r_{m(2l-1)}}{\lambda'}\cos(\theta-\phi_{m(2l-1)})} \epsilon^{-P_{m(2l-1)}} \tag{211}$$

$$v_\theta = \sum_{m=1_{(2l=2)}}^{m=q} \epsilon^{-j\frac{2\pi r_{m(2l-1)}}{\lambda'}\cos(\theta-\phi_{m(2l-1)})} \epsilon^{-P_{m(2l-1)}} \tag{212}$$

$$w_\theta = \sum_{2l=2_{(m=1)}}^{2l=p} \epsilon^{-j\frac{2\pi r_{m(2l)}}{\lambda'}\cos(\theta-\phi_{m(2l)})} \epsilon^{-P_{m(2l)}} \tag{213}$$

$$y_\theta = \sum_{m=1_{(2l=2)}}^{m=q} \epsilon^{-j\frac{2\pi r_{m(2l)}}{\lambda'}\cos(\theta-\phi_{m(2l)})} \epsilon^{-P_{m(2l)}} \tag{214}$$

The expression for the total current may be written

$$J_\theta = I_\theta u_\theta v_\theta + I_{\theta-\pi} w_\theta y_\theta \tag{215}$$

If the transducers in the circuits from each antenna of a pair are so related that

$$P_{m(2l)} - P_{m(2l-1)} = P_c, \tag{216}$$

the expression for the total current becomes

$$J_\theta = u_\theta v_\theta \left[I_\theta + I_{\theta-\pi} \epsilon^{-j[2\pi s/\lambda']\cos\theta} \epsilon^{-P_c} \right] \epsilon^{-P_1} \tag{217}$$

The directional diagram in terms of relative directional receptivity is

$$RDR = \frac{J_\theta}{J_0} = \frac{u_\theta}{u_0} \times \frac{v_\theta}{v_0} \times \left[\frac{I_\theta + I_{\theta-\pi} \epsilon^{-j[2\pi s/\lambda'] \cos\theta} \epsilon^{-P_c}}{I_0 + I_{-\pi} \epsilon^{-j[2\pi s/\lambda']} \epsilon^{-P_c}} \right] \tag{218}$$

Since there has been no assumption to this point of the character of I_θ, the significance of the coefficients u_θ and v_θ may be determined by assuming

(1) $I_\theta = I_0$, which is the directional characteristic of a vertical antenna

(2) $s = 0$

(3) $P_c = \infty$

Consideration of (218) in light of (211) and (212) under these conditions leads to the conclusion that

$$\frac{u_\theta}{u_0} \text{ and } \frac{v_\theta}{v_0}$$

are the relative directional receptivities of two arrays of vertical antennas placed at the initial ends of the antennas comprising the desired array. If, then, we designate the relationship between antennas indicated by the expression

$$J_c = \left[I_\theta + I_{\theta-\pi} \epsilon^{-j[2\pi s/\lambda'] \cos\theta} \epsilon^{-P_c} \right] \tag{219}$$

as *compensation*[13] and recognize that this expression gives the directional characteristic of a compensated antenna, we may formulate the rule that the directional characteristic of an array of similar parallel unit antennas is equal to the product of the directional characteristic of the unit antenna and the directional characteristic of an array of unit vertical antennas placed at the initial ends of the unit antennas forming the array, the product being taken point-for-point as the angle of incidence increases. The relative directional receptivity of each fundamental array of vertical antennas is termed the array factor, so that similarly, the relative directional receptivity of an array of similar parallel unit antennas is given by the product of the relative directional receptivity of the unit antenna and the array factor. This method may be extended to the solution of a complicated array such as that shown in Fig. 19, by determining the relative directional receptivity for groups of unit antennas, then determining the array factor for these groups taken as unit antennas. Expressed literally for a complex array of this type:

$$RDR_{\text{array}} = [A_1 \times A_2 \times \cdots \times A_n] RDR_{\text{unit antenna}} \tag{220}$$

where $A_1 \cdots, A_n$ are the array factors for the fundamental groups into which the complete array may be divided.

Appendix 3

WAVE TILT AND GROUND CONDUCTIVITY

In Zenneck's[31,32] exposition of the relation between the horizontal and vertical components of a plane electric wave propagated along a horizontal surface between two media, it is demonstrated that these two constituents of the wave in the upper medium (1) are related by the expression

$$-\frac{F'}{E'} = \epsilon^{j\delta} \tan T = \sqrt{\frac{\dfrac{9 \times 10^{11}}{\rho_1} + j\dfrac{1}{4\pi}\omega k_1}{\dfrac{9 \times 10^{11}}{\rho_2} + j\dfrac{1}{4\pi}\omega k_2}} \qquad (301)$$

where

SYMBOL	DEFINITION	UNIT
F'	The horizontal component of the electric wave in medium 1. (Positive direction in the direction of propagation along the interface.)	volts per kilometer
E'	The vertical component of the electric wave in medium 1. (Positive direction downward.)	volts per kilometer
ρ_1	Specific resistivity of medium 1.	ohms per centimeter cube
ρ_2	Specific resistivity of medium 2.	ohms per centimeter cube
k_1	Dielectric constant of medium 1 and equal to unity for a vacuum.	numeric
k_2	Dielectric constant of medium 2.	numeric
f	Frequency.	cycles per second
ω	$2\pi f$	

Our primary interest is in the case where the first medium is air, and the second medium is the earth beneath an antenna system. In this case the constants of the media may be given the values:

$\rho_1 = \infty$ (air)
$\rho_2 = \rho$ (earth)
$k_1 = 1$ (air)
$k_2 = k$ (earth)

Substituting these values into the general equation (301)

$$-\frac{F'}{E'}=\epsilon^{j\delta}\tan T=\frac{1}{\sqrt{k}}\left[\frac{\left(\frac{fk\rho}{18\times10^{11}}\right)^2}{1+\left(\frac{fk\rho}{18\times10^{11}}\right)^2}\right]^{1/4}\epsilon^{j[1/2\tan^{-1}(18\times10^{11}/fk\rho)]} \tag{302}$$

At this point it is desirable to indicate the significance of the term "quasi-tilt angle" as applied to T. It is seen that ($\tan T$) is the absolute magnitude of the ratio of the horizontal and vertical components of the electric field. In the case that the time phase between the two components of the field is zero (i.e., $\delta=0$), T would represent the angle of forward inclination of the propagated wave front. In general, δ is unequal to zero and hence the angle of inclination of the major axis of the ellipse traced by the electric vector is less than T, but it still remains convenient to express the ratio of the magnitudes of the two components of the field as the tangent of an angle. This angle cannot be called the wave tilt, however, but the term "quasi-tilt angle" may safely be applied to it.

The ground constants may be determined from measurement of the "quasi-tilt angle" as the following development shows:

Equation (302) may be written as two equations

$$\tan T=\frac{1}{\sqrt{k}}\left[\frac{\left(\frac{fk\rho}{18\times10^{11}}\right)^2}{1+\left(\frac{fk\rho}{18\times10^{11}}\right)^2}\right]^{1/4} \tag{303}$$

$$\delta=\frac{1}{2}\tan^{-1}\left(\frac{18\times10^{11}}{fk\rho}\right) \tag{304}$$

Solving equations (303) and (304) as simultaneous equations for δ in terms of k and T and for ρ in terms of f, k, and T yields

$$\delta=\frac{1}{2}\cos^{-1}(k\tan^2 T) \tag{305}$$

$$\rho=\frac{18\times10^{11}}{f}\,\frac{\tan^2 T}{\sqrt{1-k^2\tan^4 T}} \tag{306}$$

These two expressions have been evaluated for the extreme range of values of k that would be met in practice (k between 1 and 100) and for values of T between 0.002 and 0.2 radian and are plotted in Fig. 20. The figures for dielectric constant given by Fleming[33] show that for earth, the maximum value of k to be expected is below 20. It is evident, therefore, that δ is negligibly different from $\pi/4$ for values of T below 0.05 radian in the vicinity of an antenna which is constructed over land. Also Fig. 20 shows that the specific resistivity is practically independent of k for the same range of T. Fortunately, the measured values of T lie within these limits, so that the time phase difference between the horizontal and vertical components of an electric wave and the ground resistivity may be evaluated with but slight error from measurements of the quasi-tilt angle.

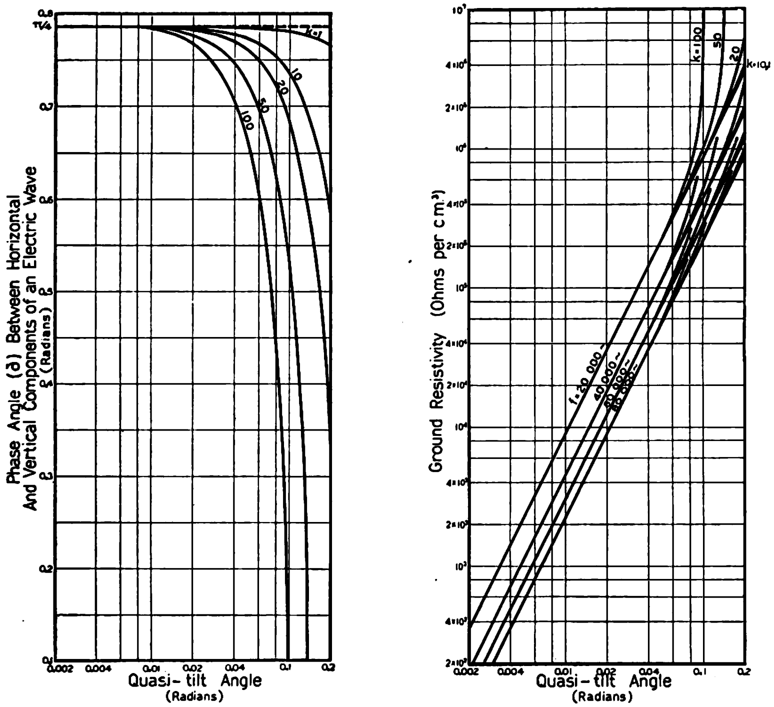

Fig. 20—Relation between Quasi-Tilt Angle, Ground Resistivity, and Phase Angle between Horizontal and Vertical Components of an Electric Wave. (By Zenneck's Formula.)

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[33] J. A. Fleming, "Principles of Electric Wave Telegraphy and Telephony," Longmans, Green and Co., 1916. 3rd edition, p. 800.

Appendix 4

PROBABILITY OF VOLTAGES GREATER THAN ANY SPECIFIED VALUE RESULTING FROM THE SIMULTANEOUS RECEPTION OF SEVERAL RADIO-TELEGRAPH STATIONS IN A RESTRICTED FREQUENCY RANGE

In order to determine the required load capacity of vacuum tubes for a radio receiver, it is necessary to obtain some estimate of the voltages from interfering signals which may occur at the input of the radio receiver and during how much of the time certain specified voltages are exceeded.

If we assume that there are N telegraph stations within a restricted frequency range, that each station contributes equal unit voltage at the receiver, and that the probability of the key being closed at any one station is constant, then the probability that exactly n stations have their keys depressed at the same time is

$$P_n = \frac{N!}{n!(N-n)!} K^n (1-K)^{(N-n)} \tag{401}$$

where K is the fraction of the total time that each station has its key depressed.

In order to determine the probability that n stations will produce a voltage equal to or greater than any specified value x we have followed Rayleigh's problem of random phases as explained in Volume 6 of his "Scientific Papers," page 618. While the conditions are not all satisfied it can be shown that they are approximately satisfied for the great majority of possible combinations and for small time intervals. The formula of Rayleigh gives the probability that the resultant of n vectors lies within an arbitrary interval $(r-dr/2,\ r+dr/2)$. Since we will assume sinusoidal voltages in the actual problem under consideration we require the probability that the projection of the resultant on the real axis is greater than a given value of x. This can be calculated by changing the polar coordinates of Rayleigh's formula to rectangular coordinates and integrating with respect to y from $-\infty$ to $+\infty$ and then with respect to x from x to $+\infty$.

The integrated formula then becomes

Probability of a voltage greater than $x = P_x$

$$P_x=A_1\Gamma\left(\frac{1}{2},\ \frac{x^2}{n}\right)+A_2\Gamma\left(\frac{3}{2},\ \frac{x^2}{n}\right)+A_3\Gamma\left(\frac{5}{2},\ \frac{x^2}{n}\right)+A_4\Gamma\left(\frac{7}{2},\ \frac{x^2}{n}\right)+A_5\Gamma\left(\frac{9}{2},\ \frac{x^2}{n}\right) \tag{402}$$

where

$$A_1=\frac{1}{2\sqrt{\pi}}\left(1-\frac{3}{16n}-\frac{5}{24n^2}+\frac{105}{16.32n^2}\right)$$

$$A_2=\frac{1}{2n\sqrt{\pi}}\left(\frac{3}{4}-\frac{25}{64n}\right)$$

$$A_3=\frac{1}{2n\sqrt{\pi}}\left(-\frac{1}{4}+\frac{155}{192n}\right)$$

$$A_4=\frac{1}{2n\sqrt{\pi}}\left(-\frac{47}{144n}\right)$$

$$A_5=\frac{1}{2n\sqrt{\pi}}\left(\frac{1}{32n}\right)$$

and

$$\Gamma(p,u\sqrt{p})=\Gamma(p)\left[1-I(u,p-1)\right]$$

in which

$$p=\frac{1}{2},\ \frac{3}{2},\ \frac{5}{2},\ \frac{7}{2},\ \frac{9}{2} \text{ and } u\sqrt{p}=\frac{x^2}{n}$$

Having found u, the I functions of $(u,\ p-1)$ can be obtained from Pearson's "Tables of the Incomplete Γ-Functions." $\Gamma(p)$ for the values of p given above is found to be

$$\sqrt{\pi},\ \frac{1}{2}\sqrt{\pi},\ \frac{3}{2}\cdot\frac{1}{2}\cdot\sqrt{\pi},\ \frac{5}{2}\cdot\frac{3}{2}\cdot\frac{1}{2}\cdot\sqrt{\pi},\ \frac{7}{2}\cdot\frac{5}{2}\cdot\frac{3}{2}\cdot\frac{1}{2}\cdot\sqrt{\pi}$$

The probability of exactly n stations being on at the same time multiplied by the probability that exactly n stations will give a voltage equal to or greater than x equals the probability of obtaining a voltage equal to or greater than x from just n stations.

Hence the summation from $n=1$ to $n=N-1$ of these probabilities for a given value of x will give the probability of obtaining a voltage equal to or greater than x from all of the N stations in the restricted frequency range considered or

$$P_{xN} = \sum_{n=1}^{n=N-1} P_n P_x \tag{403}$$

In a vacuum tube large negative voltages are equally as important as large positive voltages in producing distortion. Equation (403) has been derived for positive values greater than x but negative values greater than $-x$ are equally probable and therefore the fraction of the time that the absolute value of voltage is equal to or greater than x is $2P_{xN}$ or

$$P_{|x|N} = 2P_{xN} \tag{404}$$

Specific cases which approximate the existing conditions of long-wave transatlantic reception have been calculated from equation (404) and are shown in Fig. 21. These curves are based on the following assumptions:

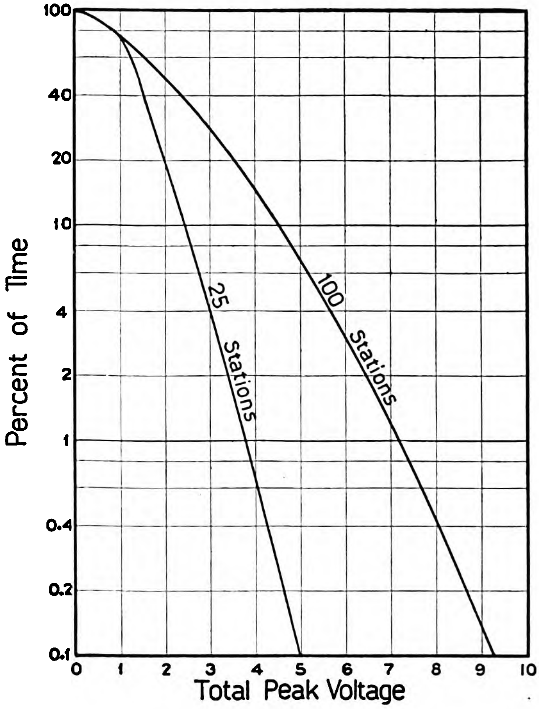

Fig. 21—Voltages Resulting from Several Unit Voltages Each Applied 15 per cent of the Time and with Random Phase and Frequency.

1. That the number of stations lying in the restricted frequency range is $N=100$ and $N=25$.
2. That each station contributes unit peak voltage to the input of the radio receiver.
3. That each station has its key depressed 15 per cent of the total time during any day. $K=0.15$.
4. That transmissions from all stations are random.

Considerable valuable assistance in the preparation of this appendix has been obtained from Dr. F. H. Murray of this department and the authors wish to express their appreciation of this aid.

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ON THE VARIATION OF GENERATED FREQUENCY OF A TRIODE OSCILLATOR DUE TO CHANGES IN FILAMENT CURRENT, GRID VOLTAGE, PLATE VOLTAGE, OR EXTERNAL RESISTANCE*

By

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Summary—*The general expressions for the generated frequency of the grid-tuned and the plate-tuned oscillators are developed. In developing these expressions it is assumed that the grid takes a convection current and that there are external resistances in the circuits. The equations which represent the frequency of oscillation of the two types of oscillators are similar and indicate that in order to transform from one type of generator to the other it is only necessary to interchange the plate-circuit constants with the grid-circuit constants.*

The effect of making any change in the circuit conditions which causes the grid current to increase is observed to cause the frequency of oscillation to decrease; and likewise any change which does not affect the grid current does not affect the frequency.

For fixed grid-battery voltage and plate-battery voltage the effect of decreasing the filament current below its rated value is to increase the generated frequency for both types of oscillators. This change in frequency is greatest for positive values of grid-battery voltage and low plate-battery voltage. With fixed filament current and plate-battery voltage, the effect of changing the grid-battery voltage from negative to positive values is to cause at first a decrease and then an increase in the generated frequency. This change in frequency is greatest for low plate-battery voltages and high (near rated) filament currents.

For fixed filament current and grid-battery voltage the effect of increasing the plate-battery voltage is first to lower the generated frequency and then to raise it.

For both types of oscillators, the effect of inserting a resistance (0 to 600 ohms) in series with the condenser is to increase the generated frequency for all values of E_c, E_b *and* I_f. *The presence of this resistance does not greatly affect the shape of the frequency-grid voltage curves. For low values of plate-battery voltage the effect of a resistance in series with the tuned coil is to cause the frequency of oscillations to increase for both types of oscillators. However, for high values of plate-battery voltage the effect of this resistance is opposite for the two oscillators. For high* E_b *the effect of increasing the resistance in the inductance branch is to decrease the frequency in the case of a grid-tuned oscillator and to increase the frequency in the case of a plate-tuned oscillator. The equations developed in this paper indicate that the grid current is responsible for the decrease in frequency with increase of the resistance in the inductance branch in the case of a grid-tuned oscillator with high plate voltage.*

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* Original Manuscript Received by the Institute, July 14, 1928. Contribution from Mendenhall Laboratory of Ohio State University.

For both types of oscillators, the effect of inserting a resistance in series with the tuned circuit as a whole is observed to cause the frequency at first to increase rapidly and then to become constant. For high values of this resistance (10,000 ohms) the frequency of oscillation is practically independent of grid voltage and particularly so for high plate voltage. For low values of plate-battery voltage the effect of a resistance in series with the exciting or tickler coil is to cause the frequency to increase rapidly for values of this resistance less than about 10,000 ohms and to cause only a slight further increase in frequency for values of resistance greater than 10,000 ohms. However, for high values of plate-battery voltage the effect of a resistance in series with the tickler coil is to decrease at first the frequency and then to increase it. The effect of this resistance is in general the same for both types of oscillators. The frequency variation is greatest for positive values of grid-battery voltage.

The grid-tuned generator will oscillate over a much wider range of variables than will the plate-tuned oscillator and frequency variation for the same change in circuit conditions will in general be less for the grid-tuned than for the plate-tuned oscillator.

The above discussion applies to oscillators without a grid-leak and grid condenser.

With properly selected values of grid-leak resistance and grid capacity both the grid-tuned and the plate-tuned oscillators may be made to generate a frequency which is very nearly that given by the equation $f = \frac{1}{2\pi\sqrt{LC}}$ *for relatively large changes in* E_c, E_b, *and* I_f. *In this work, by using a grid capacity of 0.025* μf *and a grid leak of 0.5 megohm the frequency variations were about 0.1 per cent for a 60 per cent change in plate-battery voltage and 0.08 per cent for a 30 per cent change in filament current for the grid-tuned oscillator. For the plate-tuned oscillator the frequency variations were 0.087 per cent and 0.027 per cent for 60 per cent change in* E_b *and 30 per cent change in* I_f, *respectively. These figures represent experimental results obtained near 1000 cycles per second. The theoretical equations developed seem to explain all of the experimental curves obtained.*

The results of this work indicate that with the given methods of operation, and by holding the values of the grid voltage, plate voltage and filament current constant within practical limits, the plate-tuned oscillator can readily be made to generate a frequency constant to one part in 20,000. By more refined precautions the constancy can be made still greater.

INTRODUCTION

THE variation of frequency of a triode oscillator has been noted and examined by various investigators. The work of Martyn[1] and of Eccles and Vincent[2] is of particular interest.[3] The investigation by Eccles and Vincent was carried on at radio frequencies and the results were almost entirely experimental. Martyn's work, published since the author's investigation was completed, was conducted at frequencies near 1000 cycles per second for a plate-tuned oscillator. Theoretical equations were developed and used to check the experimental observations as to order of magnitude of frequency variation.

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[1] *Phil. Mag.* 4, Nov. 1927; p. 922.

[2] *Proc.* Roy. Soc., 96 and 97 (1920).

[3] The subject of stability of thermionic oscillators at radio frequency is dealt with in a somewhat different manner by Lieut. Col. Edgeworth, *Jour.* I.E.E., 64, p. 349.

This paper is the result of a more complete study of the dependence of the generated frequency of a triode oscillator on the filament current, grid voltage, and plate voltage for two types of circuits—the grid-tuned and the plate-tuned oscillators. In addition, the effect of including external resistance in the various circuits of each type oscillator is shown. Theoretical expressions for the frequency of these two types of oscillators are developed for the case where the grid is assumed to take a convection current and where a resistance is placed in each of the external circuits. Optimum conditions of operation are determined under which the tube oscillator maintains a high degree of stability under changing conditions of filament-current plate voltage and grid voltage. This work is then confirmed experimentally for frequencies near 1000 cycles.

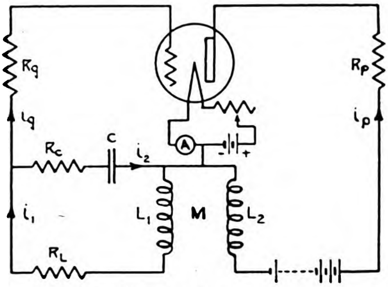

Fig. 1—Tuned-Grid Oscillator.

EXPERIMENTAL PROCEDURE

Figs. 1 and 2 show the circuits under test. In the experimental work, four single-pole double-throw knife switches were so arranged as to permit a quick change from one type of oscillator to the other. It will be noticed that coil L_1 was used as the tuned coil for both the grid-tuned and the plate-tuned oscillators. Both types of oscillators were studied at the same time. That is, for a given E_c (grid-battery voltage) and E_b (plate-battery voltage) the variation of frequency with filament current was observed for both types of oscillators before either the grid or plate voltage was changed. The grid-battery voltage was then changed and the same procedure repeated. After data for the desired range of grid voltages were thus obtained the plate voltage was changed and the entire procedure repeated. The data for the frequency variation as a function of the various external resistances were taken in a similar manner. For each curve shown all external resistances were zero except those mentioned. The filament current was maintained constant for all readings except those which were recorded as a function of filament current. If any change was made which caused the filament current, as read by the meter in the position indicated in Figs. 1 and 2, to vary, this current was brought back to the original value by readjusting the filament rheostat. Should the grid and plate return have been connected between the filament and ammeter instead of between the battery and ammeter all of the curves would be somewhat different in shape. The frequency generated was in the neighborhood of 1000 cycles per second.

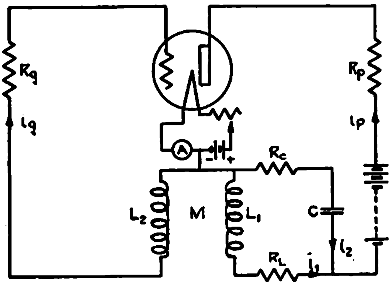

Fig. 2—Tuned-Plate Oscillator.

Various tubes of the UX-201-A type were used in these experimental tests but no difference in the general character of the frequency variation was noted. One set of coils was used throughout, and their relative positions remained fixed. The circuits used had the following constants:

$L_1 = 257.0$ millihenries
$L_2 = 213.3$ millihenries
$-M = 125.1$ millihenries
$C = 0.100$ microfarad
Resistance of $L_1 = 14.29$ ohms
Resistance of $L_2 = 12.30$ ohms

Changes in frequency of the oscillator under test were detected and measured by comparison with a standard oscillator which had been calibrated as a frequency meter. Each oscillator produced an audible tone in a telephone receiver. The frequency of the standard oscillator was set equal to that of the oscillator under test by means of audible beats.

THEORY OF TUNED GRID OSCILLATOR

In nearly all theoretical developments on the thermionic oscillator the grid current is assumed to be small enough to be neglected. In the following development the grid current is assumed large enough to be considered, a condition which frequently holds in practice.

When the grid current is not assumed to be zero it at once becomes necessary to introduce a term ν, which bears the same relation to the grid circuit as μ, the amplification factor, does to the plate circuit. Since the grid is allowed to take a convection current the expression for the grid current will be similar to that for the plate current. Both the grid and plate currents are functions of the grid and plate potentials. Therefore the fundamental relations are

$$I_p = f(E_g, E_p)$$
$$I_g = F(E_g, E_p)$$

where both E_g and E_p are variable. Suppose E_g and E_p vary in such a manner that the plate current I_p remains constant. Then

$$\frac{dI_p}{dE_g} = \frac{\partial I_p}{\partial E_g} + \frac{\partial I_p}{\partial E_p}\frac{dE_p}{dE_g} = 0.$$

If the grid current I_g is held constant,

$$\frac{dI_g}{dE_p} = \frac{\partial I_g}{\partial E_p} + \frac{\partial I_g}{\partial E_g}\frac{dE_g}{dE_p} = 0.$$

Where $E_p = (E_b + e_p)$ and E_b = plate-battery voltage
$E_g = (E_c + e_g)$ E_c = grid-battery voltage
$I_p = (I_b + i_g)$ I_b = steady plate current
$I_g = (I_c + i_g)$ I_c = steady grid current

and

$$\frac{\partial I_p}{\partial E_g} = g_m = \text{mutual plate conductance}$$

$$\frac{\partial I_p}{\partial E_p}=\frac{1}{r_p}=\text{plate conductance}$$

$$\frac{\partial I_g}{\partial E_p}=g_n=\text{mutual grid conductance}$$

$$\frac{\partial I_g}{\partial E_g}=\frac{1}{r_g}=\text{grid conductance.}$$

i_p represents the variation in I_p
e_p represents the variation in E_p
i_g represents the variation in I_g
e_g represents the variation in E_g

μ, the amplification factor, and ν, the "reflex factor"[4] are defined by the relations

$$\mu=\frac{\dfrac{\partial I_p}{\partial E_g}}{\dfrac{\partial I_p}{\partial E_p}}=-\frac{dE_p}{dE_g}=g_m r_p \text{ for } I_p \text{ held constant}$$

$$\nu=\frac{\dfrac{\partial I_g}{\partial E_g}}{\dfrac{\partial I_g}{\partial E_p}}=-\frac{dE_p}{dE_g}=\frac{1}{g_n r_g} \text{ for } I_g \text{ held constant}$$

A voltage e_g in the grid circuit is equivalent to a voltage μe_g in the plate circuit. Hence we may write

$$i_p r_p=e_p+\mu e_g$$

A voltage e_p in the plate circuit is equivalent to a voltage e_p/ν in the grid circuit. Therefore we have

$$i_g r_g=e_g+\frac{e_p}{\nu}$$

Where r_p = internal plate resistance
and r_g = internal grid resistance.

These relations will now be used in writing the circuit equations for the grid-tuned oscillator represented by Fig. 1.

$$i_p r_p=e_p+\mu e_g$$

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[4] Llewellyn. *Bell Sys. Tech. Journal,* V, July, 1926.

$$i_g r_g = e_g + \frac{e_p}{\nu}$$

$$e_p = -\left[L_2 D i_p + M D i_1 + R_p i_p\right]$$

$$e_g = -\left[L_1 D i_1 + M D i_p + R_L i_1 + R_g i_g\right]$$

$$= R_c i_2 + \frac{q}{C} - R_g i_g$$

These equations determine the polarity of M, i.e., M is intrinsically a negative quantity. $D = d/dt$; $D^2 = d^2/dt^2$ etc., and q is the instantaneous change on the condenser C.

By means of the above equations the following differential equation is obtained, expressing the current in the oscillating circuit as a function of the circuit constants:

$$a\frac{d^4 i_1}{dt^4} + b\frac{d^3 i_1}{dt^3} + c\frac{d^2 i_1}{dt^2} + d\frac{di_1}{dt} + e i_1 = 0$$

assuming a sinusoidal current $i = I\epsilon^{j\omega t}$ and substituting the proper derivatives gives

$$a\omega^4 - bj\omega^3 - c\omega^2 + dj\omega + e = 0 \tag{1}$$

where

$$a = \left\{ [L_1 L_2 - M^2]\left[r_g + \left(1 - \frac{\mu}{\nu}\right)(R_C + R_g)\right] C^2 R_C \right\}$$

$$b = \left\{ [L_1 L_2 - M^2]\left[r_g + \left(1 - \frac{\mu}{\nu}\right)(R_g + 2R_C)\right]\right.$$

$$+ (r_p + R_p)(r_g + R_g) L_1 C R_C + (r_p + R_p) L_1 C R_C^2$$

$$+ \left(1 - \frac{\mu}{\nu}\right)[(R_L + R_C) R_g + R_L R_C] L_2 C R_C + \left[(R_C + R_L) r_g L_2\right.$$

$$\left. - \frac{\mu}{\nu} L_1 R_p (R_g + R_C) \right] C R_C + \left(\mu M r_g + \frac{M}{\nu} r_p\right) C R_C^2 \left.\right\} C$$

$$c = \left\{ \left(1 - \frac{\mu}{\nu}\right) \{ L_1 L_2 - M^2 + 2 C R_C [(R_g + R_L) L_2 + R_p L_1]\right.$$

$$+ [R_L L_2 + R_p (L_1 + R_L C R_C)] C R_g + (R_L + R_g) R_p C^2 R_C^2 \}$$

$$+ (L_2 r_g + C R_C^2 r_p) C R_L + \left[(L_2 + \mu M) r_g + \left(L_1 + \frac{M}{\nu}\right) r_p \right] 2 C R_C$$

$$+\,[L_1+(R_L+R_C)CR_C]\,[(r_p+R_p)r_g+R_g r_p]C\Big\}$$

$$\begin{aligned}
d=\Big\{&\left(1-\frac{\mu}{\nu}\right)[(R_g+R_L)L_2+R_pL_1]+(L_2+\mu M)r_g\\
&+\left(L_1+\frac{M}{\nu}\right)r_p+(R_C+R_L)(r_g+R_g)(r_p+R_p)C\\
&+(r_p+R_p)(r_g+R_g+2R_L)CR_C\\
&-[2(R_L+R_g)CR_C+R_gCR_L]\frac{\mu}{\nu}R_p\Big\}
\end{aligned}$$

and

$$e=\left\{(r_p+R_p)(r_g+R_g+R_L)-\frac{\mu}{\nu}R_p(R_L+R_g)\right\}.$$

THEORY OF TUNED-PLATE OSCILLATOR

A similar analysis will now be made of the audion oscillator with the oscillatory circuit connected to the plate. Referring to Fig. 2 and using the same definitions of μ and ν, we obtain the following fundamental equations:

$$\begin{aligned}
i_pr_p&=e_p+\mu e_g\\
i_gr_g&=e_g+\frac{e_p}{\nu}\\
e_p&=-[L_1Di_1+MDi_g+R_Li_1+R_pi_p]\\
&=-\left[R_Ci_2+\frac{q}{C}+R_pi_p\right]\\
e_g&=-[L_2Di_g+MDi_1+R_gi_g]
\end{aligned}$$

As in the case of the grid-tuned oscillator, the following expression is obtained for the oscillating current:

$$a'\omega^4-b'\omega^3-c'\omega^2+d'\omega+e'=0 \qquad (2)$$

Where a', b', c', d', and e' are obtained from the expressions for a, b, c, d, and e, respectively, by changing

r_g to r_p and r_p to r_g

R_g to R_p and R_p to R_g

$\frac{1}{\nu}$ to μ and μ to $\frac{1}{\nu}$

$$\nu \text{ to } \frac{1}{\mu} \text{ and } \frac{1}{\mu} \text{ to } \nu$$

The real and imaginary parts of 1 (or 2) must vanish separately; that is,

$$a\omega^4 - c\omega^2 + e = 0 \tag{3}$$

$$-bj\omega^3 + dj\omega = 0 \tag{4}$$

and
from (4)

$$\omega^2 = (2\pi f)^2 = \frac{d}{b} \tag{5}$$

from (3)

$$\omega^2 = \frac{c \pm \sqrt{c^2 - 4ae}}{2a} \tag{6}$$

Equating (5) and (6) gives the limiting condition for oscillation to be

$$2ad - b(c \pm \sqrt{c^2 - 4ae}) = 0 \tag{7}$$

Equation (3), however, gives the expression for the frequency of oscillation. To simplify the discussion of (3) certain factors will be considered negligibly small and the variation due to the remaining factors will be examined.

THEORETICAL AND EXPERIMENTAL RESULTS

The four variables r_p, r_g, μ and ν were measured, using a General Radio hummer as a source of power. Due to the fact that the filament current was maintained constant, the form of the curves showing the variation of these variables with grid voltage is somewhat different from that usually obtained. The general shape of these curves is indicated in Fig. 3. When computing the frequency by the above formulas it is assumed that the grid potential variation, e_g, is small; in other words, that the average value over the cycle of these variables is the same as that measured by the dynamic method. If the grid potential variation e_g is not small the curve for r_g, r_p, μ or ν, as a function of grid voltage, will not necessarily be symmetrical about the points for which the calculations are made. As a matter of fact, e_g, in this work, was not small and, therefore, the values of r_g,

r_p, μ and ν, as shown by the curves in Fig. 3, are not the average values over the cycle. However, the frequency curves, as computed by the foregoing equations in which these dynamic values were used, should agree in general shape but not necessarily in numerical value, with the corresponding observed curves. The theoretical curves represent the frequency variation which would be caused by making the given change in the circuit conditions, but at the same time keeping e_g small. In other words, the computed curves assume that the magnitude of e_g is independent of any circuit change and also that e_g is small. The difference between the calculated and observed curves will, therefore, show the effect of a large alternating grid potential upon the frequency. In computing the frequency curves, μ is positive, ν is negative (in general), and M is negative.

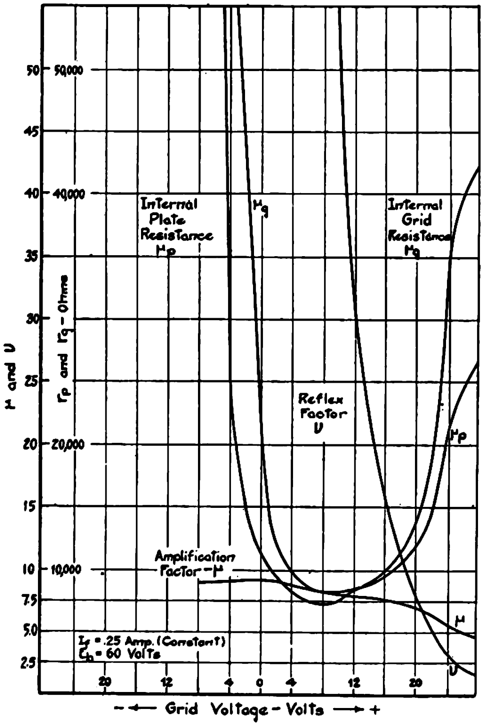

Fig. 3—Variation of Tube Parameters with Grid-Battery Voltage as Measured by Dynamic Method.

The theoretical curves assume that the tube characteristics are linear. Should the non-linear character of the tube characteristics be taken into account as in the method of Appleton and van der Pol,[5] the computed curves might be somewhat different.

VARIATION OF FREQUENCY WITH GRID-BATTERY VOLTAGE

Referring to Figs. 1 and 2 it will be noticed that when the external resistances are made zero, we have for the grid-tuned oscillator $R_c=0$ (since a low-loss condenser was used), $R_g=0$, $R_L=$resistance of L_1 and $R_p=$resistance of L_2, and for the tuned-plate oscillator $R_c=0$, $R_g=$resistance of L_2, $R_L=$resistance of L_1 and $R_p=0$.

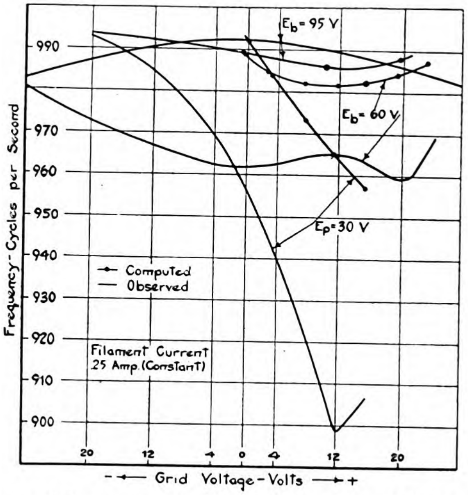

Fig. 4—Variation of Frequency with Grid-Battery Voltage—Grid-Tuned Oscillator.

Under these conditions the expression for the generated frequency of a grid-tuned oscillator becomes

$$f_{(T.G.)}=\frac{1}{2\pi}\sqrt{\frac{1+\frac{R_L}{r_g}+\frac{R_p}{r_p}+\left(1-\frac{\mu}{\nu}\right)\frac{R_pR_L}{r_gr_p}}{C\left[L_1+\frac{\beta}{r_p}\right]+\alpha\left(1-\frac{\mu}{\nu}\right)\frac{1}{r_gr_p}}} \qquad (8)$$

and making R_c and $R_p=0$ in (2) gives for the plate-tuned oscillator,

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[5] *Phil Mag.*, 43, 1922.

$$f_{(T.P.)}=\frac{1}{2\pi}\sqrt{\frac{1+\frac{R_L}{r_p}+\frac{R_g}{r_g}+\left(1-\frac{\mu}{\nu}\right)\frac{R_gR_L}{r_gr_p}}{C\left[L_1+\frac{\beta'}{r_g}\right]+\alpha\left(1-\frac{\mu}{\nu}\right)\frac{1}{r_gr_p}}} \tag{9}$$

where $\alpha=(L_1L_2-M^2)$, $\beta=(L_1R_p+L_2R_L)$, $\beta'=(L_1R_g+L_2R_L)$ and $\beta=\beta'$. Since R_L, R_p and R_g are small (about 13 ohms) compared to r_p and r_g, equations (8) and (9) reduce to

$$f_{(T.G.)}=\frac{1}{2\pi\sqrt{C\left(L_1+\frac{\beta}{r_p}\right)+\alpha\left(1-\frac{\mu}{\nu}\right)\frac{1}{r_gr_p}}} \tag{10}$$

and

$$f_{(T.P.)}=\frac{1}{2\pi\sqrt{C\left(L_1+\frac{\beta'}{r_g}\right)+\alpha\left(1-\frac{\mu}{\nu}\right)\frac{1}{r_gr_p}}} \tag{11}$$

Curves showing the variation of frequency with grid voltage under these conditions are shown in Fig. 4. These curves were obtained with a grid-tuned oscillator, and corresponding curves for the plate-tuned oscillator showed the same general characteristics as would be expected from (10) and (11). For both types of oscillator, as the plate voltage increases any given curve shows less change in frequency for the same change in grid voltage. For low values of filament current the frequency was found to be more nearly constant. Since the internal grid and plate resistances, as measured by the dynamic method, are nearly the same the computed curves for the two types of oscillators would differ very little. The *average* values of the four variables (average over one cycle) r_p, r_g, μ and ν, however, are not the same for the two oscillators, because the magnitude of e_g is not the same in the two cases. This accounts for a small difference observed in the curves for the two oscillators under these conditions.

VARIATION OF FREQUENCY WITH FILAMENT CURRENT

The effect of filament current on the frequency of oscillation can also be studied from (10) and (11). From these equations the frequency variation might be expected to be of the same general character for both oscillators, and this was found to be true. A few representative curves for the grid-tuned oscillator are shown in Fig. 5. For negative grid-battery voltage the frequency was found to be more nearly constant for low values of plate voltage. However, for positive grid voltage the frequency was more nearly constant when high values of plate voltage were used. The grid-tuned oscillator in general showed less frequency variation with filament current than did the plate-tuned oscillator.

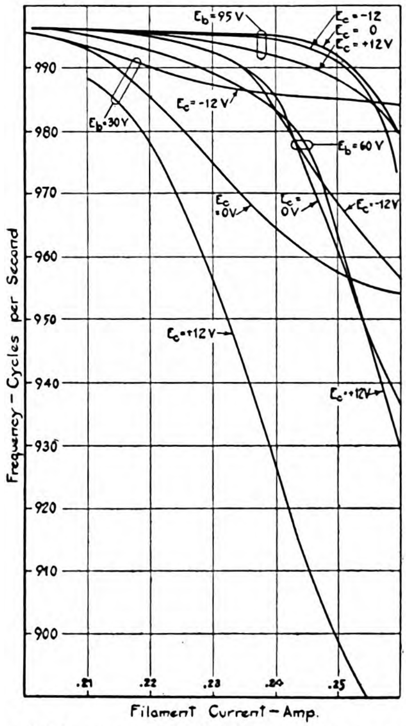

Fig. 5—Variation of Frequency with Filament Current for Various Grid and Plate-Battery Voltages—Grid-Tuned Oscillator.

VARIATION OF FREQUENCY WITH PLATE-BATTERY VOLTAGE

Figs. 6 and 7 are of interest because in addition to showing the frequency variation with plate voltage these two sets of curves are indicative of the relative range of variables over which the two types of generators will oscillate.

The curves shown in Fig. 2 of Martyn's work agree with the curves of Fig. 5 of this work when a negative grid bias is used.

However, as the grid bias is made positive the frequency is more nearly constant for the higher plate voltage.

VARIATION OF FREQUENCY WITH RESISTANCE IN SERIES WITH CONDENSER C

Assume in this case $R_g=0$, R_p=resistance of L_2, R_L=resistance of L_1 and R_c=variable. Equation (3) for the grid oscillator then reduces to (neglecting terms involving $1/r_p$ and $1/r_g$ in the numerator)[6]:

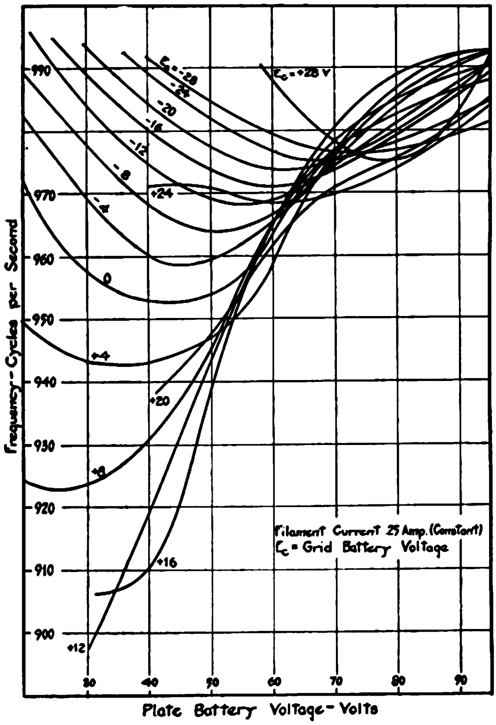

Fig. 6—Variation of Frequency with Plate-Battery Voltage for Various Grid-Battery Voltages—Grid-Tuned Oscillator.

$$f_{(T.G.)}=\frac{1}{2\pi\sqrt{C\left(L_1+\frac{\beta}{r_p}\right)+\alpha\left(1-\frac{\mu}{\nu}\right)\frac{1}{r_pr_g}+\left[\left(L_1+\frac{M}{\nu}\right)\frac{1}{r_g}+(L_2+\mu M)\frac{1}{r_p}+\beta\left(1-\frac{\mu}{\nu}\right)\frac{1}{r_gr_p}\right]CR_C}}\tag{12}$$

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[6] Equations (12) and (13) were obtained by making $R_g=0$ in the equations preceding (1) and (2), respectively.

The corresponding frequency for a plate-tuned oscillator is given by

$$f_{(T.P.)}=\frac{1}{2\pi\sqrt{C\left(L_1+\frac{\beta}{r_g}\right)+\alpha\left(1-\frac{\mu}{\nu}\right)\frac{1}{r_gr_p}+\left[(L_1+\mu M)\frac{1}{r_p}+\left(L_2+\frac{M}{\nu}\right)\frac{1}{r_g}+\beta\left(1-\frac{\mu}{\nu}\right)\frac{1}{r_gr_p}\right]CR_C}} \tag{13}$$

It is interesting to note that these two equations are the same as (10) and (11) with an additional term in the denominator.

Assume that the variables r_p, r_g, μ and ν are independent of changes in R_c. Then since R_c does not appear in the numerator, the effect of R_c will depend upon the sign of its coefficient, which coefficient depends upon the relative values of r_p, r_g, μ and ν. The term $\beta\left(1-\frac{\mu}{\nu}\right)\frac{}{r_gr_p}$ may be neglected in comparison with $(L_1+\mu M)\frac{1}{r_p}$ and $\left(L_2+\frac{M}{\nu}\right)\frac{1}{r_g}$. Therefore the sign of the coefficient of CR_c is determined by the first two terms. If the term $\left|(L_1+\mu M)\frac{1}{r_p}\right|$ is greater than $\left|\left(L_2+\frac{M}{\nu}\right)\frac{1}{r_g}\right|$ then, since M is negative, μ is positive and ν is negative (generally) the sign of the coefficient of CR_c will be negative. If L_1 is nearly equal to L_2 then the generated frequency should increase with an increase in R_c for both types of oscillators. This is shown to be true by curves *A* and *B* of Figs. 8 and 9.

VARIATION OF FREQUENCY WITH RESISTANCE R_L, IN SERIES WITH THE TUNED COIL L_1

The expressions for the generated frequency are the same as those given in (8) and (9). However, we shall now consider R_L variable. The effect of R_L may be seen a little more clearly, perhaps, if (8) and (9) are rewritten as follows:

$$f_{(T.G.)}=\frac{1}{2\pi}\sqrt{\frac{r_p+R_p+R_L\frac{r_p}{r_g}+\left(1-\frac{\mu}{\nu}\right)\frac{R_pR_L}{r_g}}{C[L_1r_p+L_1R_p+L_2R_L]+\alpha\left(1-\frac{\mu}{\nu}\right)\frac{1}{r_g}}} \tag{14}$$

and

$$f_{(T.P.)}=\frac{1}{2\pi}\sqrt{\frac{r_p+R_g\dfrac{r_p}{r_g}+R_L+\left(1-\dfrac{\mu}{\nu}\right)\dfrac{R_gR_L}{r_g}}{C\left[L_1r_p+(L_1R_g+L_2R_L)\dfrac{r_p}{r_g}+\alpha\left(1-\dfrac{\mu}{\nu}\right)\dfrac{1}{r_g}\right.}} \qquad (15)$$

Remembering that R_p in (14) is equal to R_g in (15), it is clear that if r_g is nearly equal to r_p, the effect of increasing R_L will be the same for both types of circuits. If, however, r_g is greater than r_p, and this difference is great enough, a given increase in R_L will cause the denominator of (14) to increase more than the numerator; that is, to increase R_L in (14) would decrease the generated frequency. For the plate-tuned oscillator, to increase R_L would cause a greater increase in the numerator; that is, to increase R_L in (15) would increase the generated frequency. For example, suppose the grid current is zero (r_g infinite), (14) and (15) reduce to

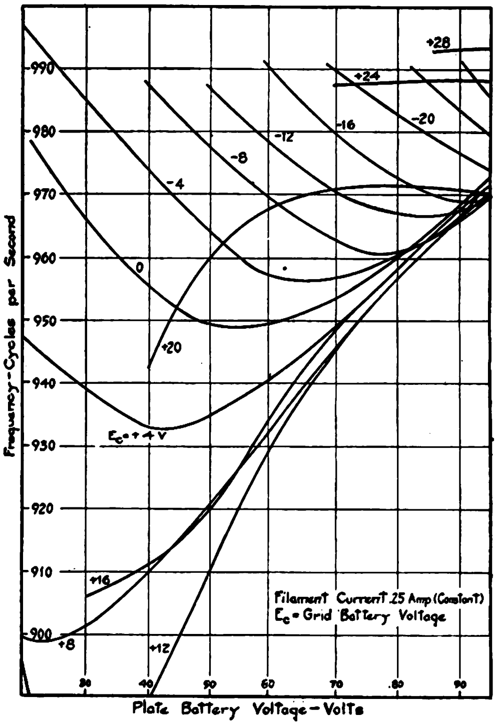

Fig. 7—Variation of Frequency with Plate-Battery Voltage for Various Grid-Battery Voltages—Plate-Tuned Oscillator.

$$f=\frac{1}{2\pi}\sqrt{\frac{1+\frac{R_p}{r_p}}{C\left[L_1+(L_1R_p+L_2R_L)\frac{1}{r_p}\right]}} \tag{16}$$

and

$$f=\frac{1}{2\pi}\sqrt{\frac{1+\frac{R_L}{r_p}}{CL_1}} \tag{17}$$

Clearly in (16) and (17) the effect of R_L is opposite in the two cases. The effect of R_L on the frequency variation for a low and high plate voltage is shown by curves C and D of Figs. 8 and 9. For low plate voltage the frequency increases with increasing R_L for both oscillators while for high plate voltage the frequency increases for the plate-tuned oscillator and decreases for the grid-tuned oscillator as R_L is increased.

THE VARIATION OF FREQUENCY WITH EXTERNAL GRID AND PLATE RESISTANCE

To study this effect, make the external coil and condenser resistance zero as before. Putting $R_c=0$ in (1) and rearranging gives for the grid oscillator,

$$\underset{(T.G.)}{f}=\frac{1}{2\pi}\sqrt{\frac{1+\frac{R_p}{r_p}+(R_g+R_L)\frac{1}{r_g}+\left(1-\frac{\mu}{\nu}\right)\left(R_L+R_g\right)\frac{R_p}{r_gr_p}}{C\left[L_1+L_1\left(\frac{R_p}{r_p}+\frac{R_g}{r_g}\right)+L_2\frac{R_L}{r_p}\right]+\left[\alpha+(L_1R_p+L_2R_L)CR_g\right]\left(1-\frac{\mu}{\nu}\right)\frac{1}{r_gr_p}}} \tag{18}$$

The corresponding equation for the plate oscillator is

$$\underset{(T.P.)}{f}=\frac{1}{2\pi}\sqrt{\frac{1+\frac{R_g}{r_g}+(R_p+R_L)\frac{1}{r_p}+\left(1-\frac{\mu}{\nu}\right)\left(R_L+R_p\right)\frac{R_g}{r_gr_p}}{C\left[L_1+L_1\left(\frac{R_g}{r_g}+\frac{R_p}{r_p}\right)+L_2\frac{R_L}{r_g}\right]+\left[\alpha+(L_1R_g+L_2R_L)CR_p\right]\left(1-\frac{\mu}{\nu}\right)\frac{1}{r_gr_p}}} \tag{19}$$

Inspection of (18) and (19) will show that, other things remaining constant, to change the external grid resistance in the plate-tuned oscillator will cause the same sort of frequency variations as to change the external plate resistance in the grid-tuned oscillator and vice versa. However, when the external plate resistance is changed the internal plate resistance does not remain constant because the plate voltage is lowered by an amount equal to the IR drop across the external resistance. For a similar reason the internal grid resistance does not remain constant when an external grid resistance is added. The internal grid resistance goes through the same kind of changes with respect to filament current, grid voltage and plate voltage as does the internal plate resistance (see Fig. 3). Therefore, we may still expect changes in R_p to cause the same kind of changes in frequency for one type of oscillator as changes in R_g do for the other type.

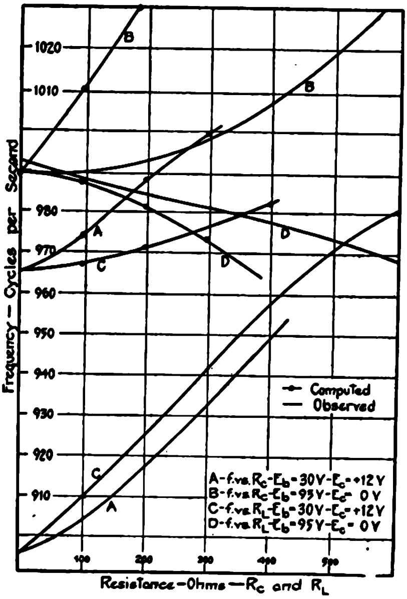

Fig. 8—Variation of Frequency with External Coil and Condenser Resistance—Grid-Tuned Oscillator.

Now suppose R_g to be made large, 20,000 ohms or more. As the external grid resistance is increased the internal grid resistance will increase because the I_gR_g drop will lower the average grid potential. The internal plate resistance will also increase but in general less rapidly than r_g. Hence, for large external grid resistance (18) and (19) both approximate the relation

$$f = \frac{1}{2\pi\sqrt{LC}}.$$

Fig. 10 shows the effect of external plate resistance on a tuned-grid oscillator and Fig. 11 shows the effect of external grid resistance on a tuned-plate oscillator. The similarity of these two sets of curves is accounted for by (18) and (19). The same similarity was observed for lower values of plate voltage.

THEORY OF THE OSCILLATOR WITH GRID LEAK AND GRID CONDENSER

The elements of the grid circuit are indicated in Fig. 12.

For a thermionic tube to produce self-sustaining oscillations there must be a definite amount of energy transferred from the plate circuit back into the grid circuit. Suppose the grid-leak resistance R_1 to be infinite, then clearly the grid will in a short time acquire sufficient negative charge to reduce the plate current and hence the available energy to zero. Under these conditions the circuit will obviously not oscillate. There is then a certain definite maximum charge which the grid condenser C_1 can acquire and yet allow the circuit to oscillate, for any given circuit conditions.

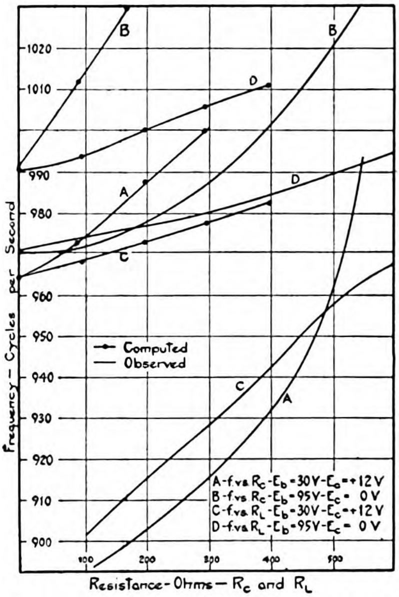

Fig. 9—Variation of Frequency with External Coil and Condenser Resistance—Plate-Tuned Oscillator.

The charge Q_1 on C_1 at any time t after the charge Q_0 starts to diminish is given by

$$Q_1 = Q_0 \epsilon^{-\frac{t}{R_1 C_1}} \quad \text{or} \quad \frac{Q_1}{Q_0} = \epsilon^{-\frac{t}{R_1 C_1}}$$

Let Q_1 be the maximum charge on C_1 which will allow the circuit to oscillate and Q_0 be the charge on C_1 when $t=0$ (or in other words, let Q_0 be the charge that would accumulate on C_1 if R_1 were infinite); Q_1 and Q_0 are both constant for given circuit conditions. The minimum time in which Q_0 must be reduced to Q_1 is equal to the period of oscillation of the circuit, otherwise a sufficiently large charge would accumulate on the grid and eventually "block" the tube. Hence we have

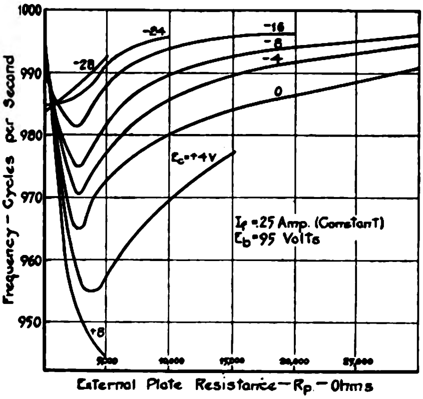

Fig. 10—Variation of Frequency with External Plate Resistance for Various Grid-Battery Voltages—Grid-Tuned Oscillator.

$$\frac{Q_1}{Q_0} = \epsilon^{-\frac{1}{fR_1C_1}} = \text{constant} < 1$$

Where f = frequency of oscillation of the circuit.

Thus $fR_1C_1 = K$. Therefore, if any two of these quantities are held constant and the third increased, the negative charge on the grid will increase and stop oscillations. Thus if f and R_1 are fixed and the capacity C_1 is increased, the negative charge on C_1 will increase and the circuit will not oscillate continuously but in beats. If R_1 is now lowered so that the charge can leak off before it reaches the critical value the circuit will again oscillate continuously. If the time constant C_1R_1 is lowered, Q_1 decreases and the grid becomes more positive. With $C_1=0$ and R_1 is finite, we have simply an external grid resistance.

A circuit using a grid leak and condenser should, therefore, be expected to generate a nearly constant frequency for relatively large changes in the circuit constants. Since the grid is maintained negative throughout most of the cycle, we may neglect any term or terms in the foregoing equations which involve $1/r_g$. Thus (8) reduces to (16). If R_p/r_p be neglected, this equation reduces to the usual expression for the frequency of a grid-tuned oscillator, namely,

$$f=\frac{1}{2\pi\sqrt{C\left(L_1+L_2\frac{R_L}{r_p}\right)}}$$

and (10) reduces to (18), which is the usual form for a plate-tuned oscillator.

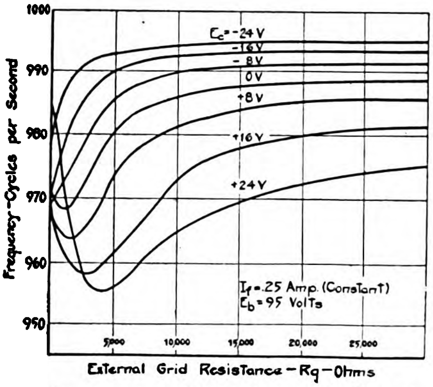

Fig. 11—Variation of Frequency with External Grid Resistance for Various Grid-Battery Voltages—Plate-Tuned Oscillator.

Careful determinations were made of the frequency variations for changes in E_c, E_b, and I_f. The most desirable size of grid leak and grid condenser depends, among other things, upon the frequency of oscillation and the input impedance of the tube. Since the impedance of the condenser must be low compared to that

TABLE I

E_b=95 volts I_f=0.25 amp.		E_b=95 volts E_c=−4 volts		I_f=0.25 amp. E_c=−4 volts	
Change in E_c volts	Per cent change in frequency	Change in I_f amp.	Per cent change in frequency	Change in E_b volts	Per cent change in frequency
Grid-Tuned Oscillator					
+30 to −30	0.0013	0.20 to 0.26	0.080	30 to 95	0.105
Plate-Tuned Oscillator					
+10 to −10	0.007	0.20 to 0.26	0.027	30 to 95	0.087

of the tube a larger condenser should be used for low frequency than for high frequency. Table I indicates the frequency variation observed while using a grid condenser of 0.025μf and a grid-leak resistance of 0.5 megohm.

This small frequency variation shows that it is possible to construct a triode oscillation generator, for low frequencies at least, which can be maintained constant. Obviously, if the generator is to be used as a standard, no such changes, as indicated in the table, will be made in E_c, E_b, or I_f.

GENERAL DISCUSSION

The average grid current and average plate current were observed along with the frequency for nearly all of the runs made. The variations of these currents are not shown, but the following facts were observed. In order that the generated frequency be nearly that which is predicted by the equation $f=\frac{1}{2\pi\sqrt{LC}}$ the grid current (as measured by a d.c. meter) must be small. It was also observed that for the generated frequency to be constant for any given change of circuit conditions, the grid current must also be constant for that same change. It will be noticed that these two conditions are both satisfied if we have a large external grid resistance, or better still, if the proper values of grid condenser and grid leak are used.

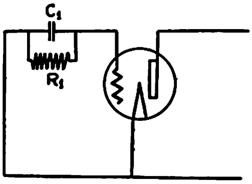

Fig. 12

It is interesting to note that in all cases, except when a grid leak and grid condenser were used, a given change in the circuit conditions caused a smaller frequency variation in the case of a grid-tuned oscillator than in the case of a plate-tuned oscillator. However, when a grid leak and grid condenser were used the plate-tuned oscillator was the more nearly constant for all changes in circuit conditions except changes in E_c.

The frequency variation was observed for several different ratios of L/C (keeping L times C constant) and it was found that this variation was less when a small L and correspondingly large C were used.

The results of this work indicate the following set of circuit conditions to be very suitable for a constant frequency triode oscillation generator, using a UV or UX 201-A tube and for a frequency range of 1000 ± 100 cycles:

Plate-battery voltage about	100 volts
Grid-battery voltage about	−4 volts
Filament current	0.25 amp. (rated)
Grid condenser	0.025 μf
Grid leak	0.5 megohm

The plate-tuned oscillator is in general more nearly constant under these conditions.

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SOUND MEASUREMENTS AND LOUDSPEAKER CHARACTERISTICS*

BY

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Summary—A brief description of the method used to measure loudspeaker response is given. The Rayleigh disk and condenser microphone are compared as sound detectors. A number of loudspeaker sound pressure response curves are shown, and interpreted in terms of pleasantness of reproduction, as determined by low- and high-frequency cut-off, smoothness of response, and tone balance. Tube overloading and the effect of loudspeaker response on its apparent accentuation or diminution is discussed. The effect of room absorption characteristics, room resonances, position of the loudspeaker in the room, and position of the listener with respect to loudspeaker on loudspeaker reproduction is explained by means of diagrams and loudspeaker response curves.

THE success of an attempt to put together the electrical and mechanical apparatus known as a radio set will depend more on the character of loudspeaker used than on any other part of the system. Since the beginning of broadcasting the sound reproducer has usually been the weakest part of the chain which leads to exact reproduction.

Very little actual data has been published on the performance of loudspeakers. The reason for this lack of information lies in the few laboratories which have been equipped to get a satisfactorily exact measure of the loudspeaker performance, and the reluctance of those who have the data to publish what they have, because of the misinterpretation of the performance which the general unfamiliarity with loudspeaker curves would lead to.

The measure of the goodness or poorness of an amplifier from the standpoint of frequency response is usually very simple. Its defects are generally a falling off at the high and low ends with a quite smooth response in the middle. The amplifier which falls off least and has the greatest frequency range is the best, and comparison is simple.

To those who are familiar with good amplifier curves, the first loudspeaker response record is quite shocking. A profile

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map of the Rocky Mountains may be quite smooth in comparison. When a curve is of this type, it requires interpretation and it is the purpose of this paper to try to interpret in terms of pleasantness of reproduction some response curves which we have taken on various loudspeakers, some commerical, others just laboratory experiments.

Loudspeaker frequency response is usually described very indefinitely. It is mellow, it is tubby, nasal, shrill, reproduces all frequencies evenly. What do these words mean in terms of sound output? When a loudspeaker is described as having fine low or high frequencies, how high or low is meant?

Let us digress for a moment to review briefly the methods by which response curves are taken.[1] The outline of procedure is as follows: (1) Deliver energy to the loudspeaker to simulate what it would get from the last tube of a perfect set. (2) Measure the sound energy at a selected point in front of the loudspeaker at all audible frequencies. The apparatus is shown schematically in Fig. 1. The first part is the same every day electrical problem which is encountered in the measurements of most radio apparatus and needs no further comment.

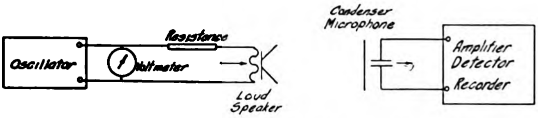

Fig. 1—Schematic Diagram of Loudspeaker Measurement System.

The second part requires an instrument for measuring sound energy. This instrument for most convenient use should be portable, relatively rugged, and as small as possible, so as not to distort the sound field, due to reflections from its surface. The familar condenser transmitter used for broadcast pickup meets the first two requirements but not the latter. The pickup used should preferably be not more than $\frac{1}{2}$ in. in diameter instead of the three inches approximately of the commerical condenser microphone. It is probably only a question of time until one small enough and at the same time sensitive enough is developed. For the present we must correct our results for the distortion caused.

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[1] "Loud Speaker Response Measurements": Erwin Meyer, *Elect. Nach. Tech.*, 3, 290, 1926; 4, 86, 1927; 4, 203, 1927. *Zeit. für Tech. Physik*, 7, 12, 1926. Cohen, Aldridge, and West, *Jour.* I. E. E., 64, 1023; 1926. E. Gerlach, *Zeit. für Tech. Physik*, 11, 1927. F. Trendelenberg, *Wiss. Veroffentlichungen aus Dem Siemens Koncern*, 5, 1926. I. Wolff and A. Ringel, Proc. I. R. E., 15, 363; May, 1927.

A sound wave is characterized by a to and fro motion of the air, and an oscillatory pressure. In a free almost plane wave, at any point in space the two are practically in phase, and the total energy is very nearly equally divided between that stored as kinetic energy of motion of the air and the potential energy of compression. None of the sound measuring devices used measures the energy directly. Since, however, the energy is equally divided between that connected with velocity and that connected with pressure either of the two may be used as the measuring stick.

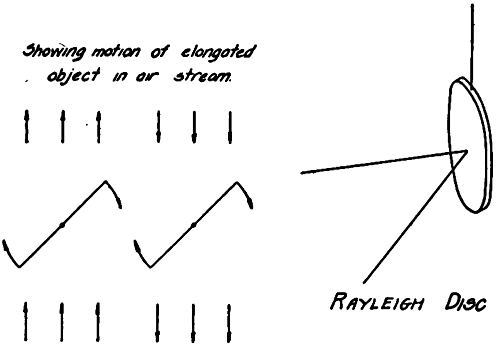

Fig. 2

The condenser microphone depending for its action on the change in capacity caused by the motion of one charged plate close to another is a pressure operated device, since the motion of the plate is caused by the force which is applied to it.

Another instrument which is used for making sound measurements and which has been most extensively used in Germany depends for its operation on the velocity component of the energy.[2] This neat little device is one of the numerous contributions of Lord Rayleigh to the science of acoustics and has been called the Rayleigh disk. As it is not as familiar as the condenser microphone I think that a rather brief description of it and its action will be interesting. It is illustrated in Fig. 2.

The Rayleigh disk depends for its action on the well known phenomenon that an elongated object pivoted at the center when placed in a flowing stream will tend to set itself with its long direction perpendicular to the lines of flow. Sound waves consisting of a to and fro motion of the air particles constitute an alternating air stream and will tend to make such an object set itself at right angles to their flow lines, or, which is the same, at right angles to the direction of propagation of the sound wave.

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[2] "Rayleigh Disc:" Erwin Meyer, *Elect. Nach. Tech.*, 3, 290, 1926. E. J. Barns and W. West, *Jour.* I. E. E., 65, 871, 1927. Charles H. Skinner, *Physical Review*, 27, 346, 1926.

The force tending to turn the object will be stronger as the intensity of the sound wave is stronger. This is quite evidently a device which depends for its action on the velocity of motion of the air particles rather than the pressure which may exist in the wave.

In practice a very thin, small, silvered, mica disk is suspended by a thin fiber. Light reflected from the silvered surface is used to measure the deflection of the disk. A knowledge of the torque brought into play by the suspension fiber and the forces caused by the air flow allow the disk to be calibrated. The measurement then consists of a balancing of these two forces.

The Rayleigh disk has the advantage that it can be made very small (within the $\frac{1}{2}$ in. limit), and therefore causes very little distortion of the sound field. On the other hand the deflecting forces brought into play by sound fields of normal strength are quite minute and require the use of a delicate fiber suspension, so that the instrument is neither stable nor portable.

The loudspeaker response curves shown in this paper have all been taken using the condenser microphone and follow quite closely the procedure which was described more in detail in an article by Mr. Ringel and myself.[3]

In interpreting loudspeaker response curves for the purpose of estimating the fidelity of the loudspeaker four points should be particularly noted:

1. Low-frequency cut-off
2. High-frequency cut-off
3. Smoothness of response between cut-off points
4. Balance between high and low frequencies.

The effects of some of these factors are fairly well known. The cut-off of low frequencies makes the reproduction lose fullness and body. The removal of high frequencies takes away crispness and distinctness, removing the sibilants from speech.

The interpretation of an uneven response is more difficult and depends on the particular kind of defect which is introduced.

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[3] I. Wolff and A. Ringel, loc. cit.

The type of distortion noticed will be described in connection with the curves which will be shown to illustrate a few of the infinite possible variations.

When a loudspeaker has certain defects such as excessive response in a part of the low-frequency region, a partial compensation may be made by introducing a similar rise in the high-frequency response. In general there is a certain frequency which may be considered as a center of gravity point and which is somewhere in the neighborhood of 1000 cycles, above and below which the total frequency response should about balance when plotted on a logarithmic or musical frequency scale. The response curves shown in Fig. 3 for two horn loudspeakers which were popular in the early stages of radio illustrate the fulfillment and violation of some of these requirements.

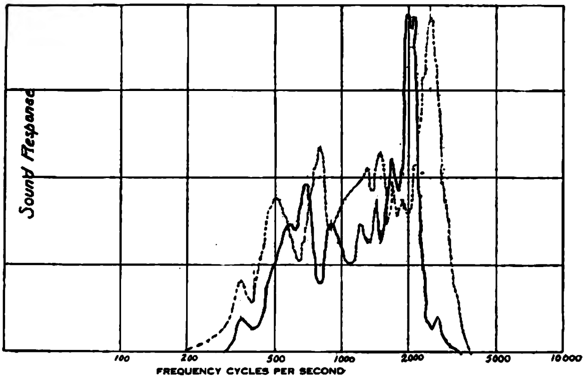


Fig. 3—Old Horn Speaker.

It is interesting to take these speakers for study since they were developed in the days before sound pressure curves were known and therefore any peculiarities are certainly the result of an attempt to adjust the materials at hand to give the most pleasing results. It is quite evident that the designers independently chose a frequency of about 1000 as a center and spread their possible frequency range as much as possible on either side. It would have been perfectly simple in either case by simply changing the diaphragm stiffness to have shifted the curves either up or down on the frequency scale, but this apparently would have made the tone balance poor.

The frequency range covered by these loudspeakers is very mediocre, and accounts for what is known as the characteristic small horn quality. This quality has usually been associated with mechanical reproduction and has been called a metallic quality. This is really a false name for it, as what is heard is only the result of the restricted frequency band, modified by whatever peaks and depressions happen to be present in the transmission range.

The next curve, Fig. 4, shows the response of a more recent loudspeaker which is typical of some of the moving iron speakers which are being sold today. For purposes of comparison the response of the better of the two horns shown in the preceding slide is again given.

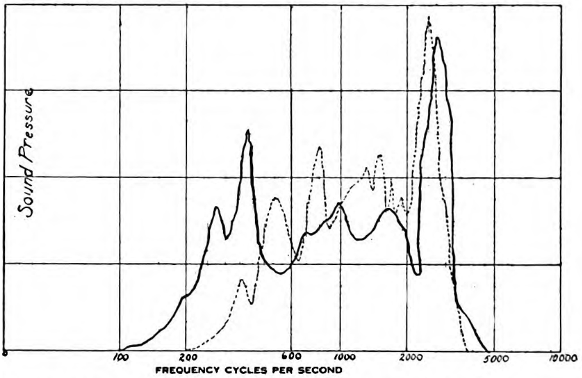


Fig. 4—Illustrating Response with Low- and High-Frequency Peaks.

Several interesting points are brought out by this curve. The low- and high-frequency cut-off points have been considerably extended, which of course means more satisfactory quality. In fact, the extension is now sufficient to remove definitely the metallic, mechanical timbre from reproduction as given by this instrument. It is again interesting to see how closely the area under this curve when plotted on a logarithmic frequency scale balances in the neighborhood of 1000 cycles. You will notice that the response has peaks at both the lower and upper ends. This is a device which is often resorted to in order to give the impression of an extension of the frequency range. The ear would notice the lack of low frequencies due to the failure to respond much below 200 cycles. The peak shown causes a series of notes in the lower musical range to be exaggerated. To the untrained ear this creates the impression of very good low-frequency response. The resonant peak type of low-frequency response is very easily detected by its booming character and by the continual loudness of certain notes in a limited musical range.

The addition of this low-frequency peak alone would unbalance the response, putting too much emphasis on the lower register; in fact, it is the equivalent from this standpoint of extending the range about one octave lower. For various reasons it may be impossible to extend the higher end to get equilibrium. The next best thing to do would be to raise the whole response above 1000 cycles. If this is still not possible a peak must be added somewhere in the upper range to give the impression of sufficient high-frequency response. The peak at about 2500 cycles shown in this curve does just this, and has the advantage of being high enough to create the impression of true high frequencies being produced.

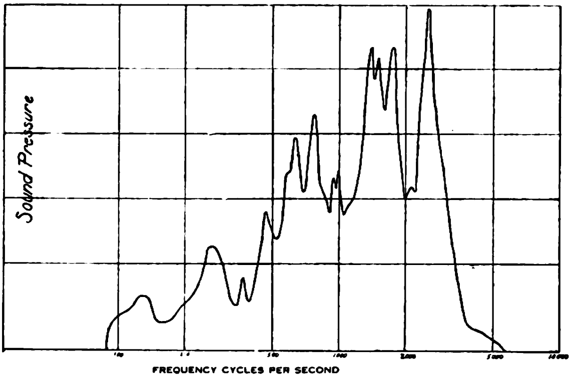


Fig. 5—Typical Loudspeaker Response.

A distinctly different kind of response, which appeals to some people, is shown in Fig. 5. This shows a more extended low-frequency response with high-frequency cut-off about the same as on the preceding curve. The balance point is again in the neighborhood of 1000 cycles. While the preceding speaker had its high parts at the two ends with a general depression in the middle, this one concentrates on the high middle range with a tapering off towards the ends. Whether one or the other is preferred is a matter of personal taste. The rising response from 500 to 2500 cycles gives a nasal quality to the reproduction which, when combined with the sharp cut-off at the latter point, makes this speaker quite successful in masking tube overloading.

Under present broadcasting conditions where the range of frequencies transmitted cuts off pretty sharply at 5000 cycles or below, tube overloading on a speaker which reproduces real high frequencies shows up as a roughness, rasp, and very often as a sound which resembles a paper rattle. This is caused by the generation of harmonics and combination tones. These added notes show up particularly badly when they are produced at the higher frequencies, as there is no true transmitted sound of the same frequency to act as a mask.

Fig. 6 illustrates the amount of cutting off which was required in a certain speaker to mitigate this overloading effect. The exact position of the cut-off is a matter of judgment. In general, the lower it is made the less the overloading distortion. However, if this is carried too far, the fidelity at the higher frequencies will be reduced.

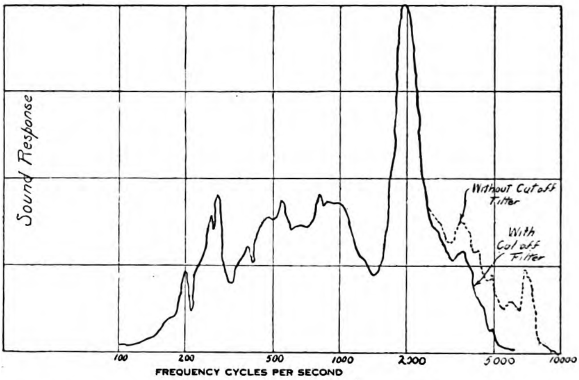


Fig. 6—Showing Elimination of Tube Overloading.

Up to this point we have been considering loudspeaker response from the standpoint of the sound radiated by the loudspeaker. We will now take up the equally important matter of the effect of external conditions on the sound heard by the listener.

We are sometimes annoyed after having conducted listening tests on a loudspeaker, and having reached the conclusion that it is pretty good, to find it unsatisfactory when moved to a different room or even a different position in the same room. It is therefore very important when taking loudspeaker curves to consider the question of room acoustics and loudspeaker position.

The response curves I have shown have all been taken in a room with absorption characteristics which were intended to simulate those encountered in the ordinary home, and the microphone has been placed so as to receive the sound which could be expected to reach the listener's ear.

The loudspeakers will, however, be used in a wide variety of rooms and in interpreting the response characteristics we should know what changes will be made when some of the conditions are made different.

Some of the factors which may be expected to have a pretty big effect are:

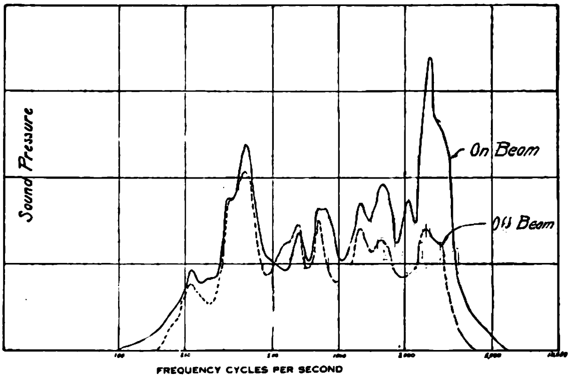
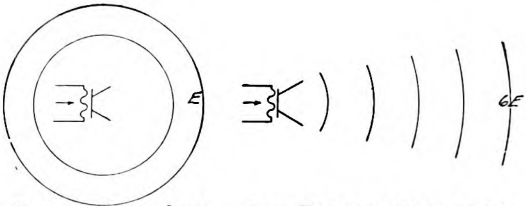


Fig. 7—Illustrating Beam Effect.

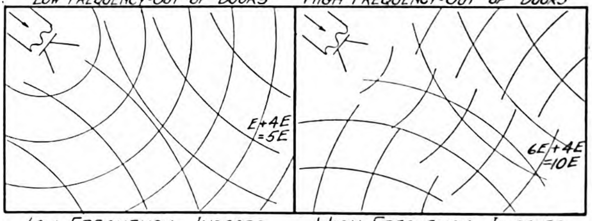
Room absorption characteristics
Room resonances
Position of loudspeaker in room
Position of listener with respect to loudspeaker.

Consider the position of listener first. Most loudspeakers radiate more sound towards the front than to the sides, particularly at the higher frequencies where the sound comes out almost in the form of a beam. This effect is a general property of wave radiation and we may generalize by saying that whenever the vibrating surface is small compared to the wavelength being emitted the sound will spread equally in all directions and that for wavelengths small compared to the surface the sound will be sent out in the form of a beam perpendicular to the surface. The result is greater high-frequency response directly in front of the speaker, as illustrated in Fig. 7.

The extent to which the beam effect modifies the loudspeaker response as heard by the audience is also a question of room absorption. To illustrate: Consider a loudspeaker which sends out the same total amount of energy at all frequencies, but concentrates the higher frequency energy so that the ratio of high-frequency to low-frequency energy radiated directly in front is six to one. The sound when it hits the walls of the room will be partially reflected, the amount of the reflection depending on the materials making up the room. In order to make the discussion as simple as possible assume that the room has the same absorption at all frequencies.



LOW FREQUENCY-OUT OF DOORS HIGH FREQUENCY-OUT OF DOORS



LOW FREQUENCY - INDOORS HIGH FREQUENCY - IN DOORS

Fig. 8—Illustrating Effect of Room on Loudspeaker Response.

The sound energy reaching the listener may be considered as made up of two parts, one due to direct radiation from the loudspeaker, the other coming from successive reflections from the walls and other objects in the room. Experiments have shown that the latter energy is pretty uniformly distributed throughout the room after several reflections, leading to the conclusion that the magnitude of the reflected energy will depend only on the total energy radiated by the loudspeaker. When the absorption is small the reflected energy reaching the listener's ear is much bigger than that received directly from the loudspeaker. In a room with small uniform absorption at all frequencies the energy reaching the listener's ear is determined mostly by the total energy radiated by the loudspeaker; in a room with high absorption it is determined mostly by the energy radiated in the direction of the listener. This leads to the conclusion, which is a little surprising at first, that the loudspeaker response is greatly modified by the room even though it has uniform frequency absorption.

Returning to the loudspeaker which radiates the same amount of energy at all frequencies, but which concentrates the higher frequencies in the form of a beam so that the energy density directly in front is six times that for the lower frequencies, some interesting numerical results may be obtained which are shown diagrammatically in Fig. 8.

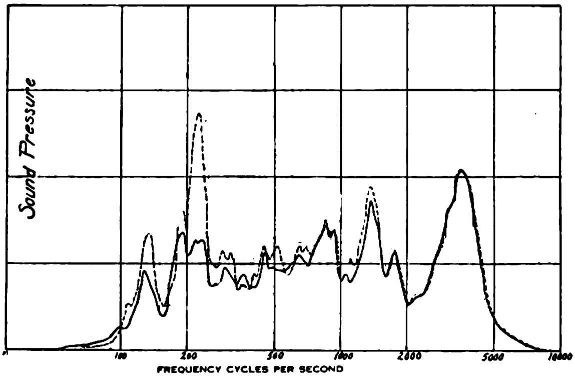


Fig. 9—Illustrating Cavity Resonance.

If the speaker is in a room with very high absorption, or better yet outdoors (equivalent to a room with absorption coefficient equal to 1), the amount of energy reaching the listener is determined entirely by the direct radiation and is in the ratio 6 units high to 1 unit low. When the loudspeaker is placed in a room of smaller absorption the reflected energy may easily be four times as big as the energy which is received by direct radiation when there is no beam. This energy is the same for both high and low frequencies, since the total energy radiated is the same, so that 4 units are added to each, making the ratio 10 units high to 5 units low or 2 to 1 instead of 6 to 1 as previously.

The preceding explains to a certain degree the dropping off of low tones which is very apparent when a loudspeaker is placed in the open. The loss is even more accentuated by the fact that the absorption of the room is not uniform at all frequencies, but is usually smaller at the low end. Obviously this leads to a greater reflected energy for low tones in the room, and a greater relative decrease in their intensity when the loudspeaker is outdoors.

I will not devote much space to the influence of frequency variations in the room absorption coefficient except to say that the effect can be very considerable. It is perfectly apparent that small absorption at any frequency leads to an increase in intensity at that frequency and that the opposite is true for high absorption.

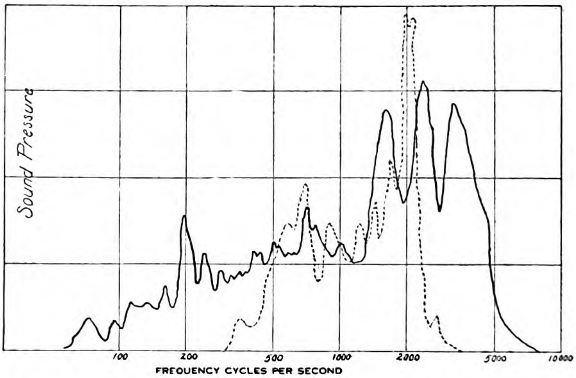


Fig. 10—Modern Loudspeaker in Comparison with Old Horn.

The air in a room which is not very absorbing may resonate to certain frequencies just like any other mechanical vibrator. If the loudspeaker is coupled to the air in the room closely enough this resonance may be excited. There are certain places in the room where the coupling will be such as to induce this resonance more vigorously.

This raises the general question of the effect of the position of the loudspeaker in the room on the response. We have already seen that high frequencies are radiated in a beam. If high response is wanted the speaker should therefore be pointed and placed so as to cover as large a portion of the audience as possible. Placing the loudspeaker in a corner or in any kind of a cavity will usually have a big effect on the response. The space between the back of the speaker and wall or other obstruction will act as a resonant chamber whose vibrations will be excited by the vibrations of the rear side of the speaker diaphragm. It is impossible to say whether this effect will be pleasing or otherwise.

It will depend on the unadulterated response characteristic and whether the resonance is of such frequency as to supply a region which is lacking. The change in the loudspeaker response caused by this resonance is illustrated in Fig. 9.

We have now seen some of the reasons why radio reproduction is still far from the ideal of being the same as the original. Fig. 10 shows a speaker of the household type which compares favorably with anything which has been produced to date and which represents fairly well the better grade of loudspeaker reproduction attainable at this time. The room left for improvement needs no comment.

In order to conclude in a little more optimistic mood the response of the old horn speaker shown in Fig. 3 is superposed for purposes of comparison. The change is at least encouraging.

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THE DESIGN OF TRANSFORMERS FOR AUDIO-FREQUENCY AMPLIFIERS WITH PREASSIGNED CHARACTERISTICS

By

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Summary—*The requirements of an ideal transformer are stated, and the difficulties encountered in attempting to build transformers for interstage coupling units which will meet these requirements are pointed out. The aim in design for audio amplifiers is stated to be a reasonable voltage amplification for one tube and one transformer which is independent of the frequency over a range necessary for broadcast reception.*

The equivalent a.c. circuit for some types of transformers with tube source and tube load is set up and solved. The impedance and voltage amplification characteristics are explained from the solution of the equivalent circuit. Expressions for calculating the constants of the transformer are given. A study of the design relations and voltage amplification characteristics reveals difficulties which are encountered in design and some methods for overcoming these difficulties. The effect of the continuous flux in the core of the transformer is illustrated and a scheme is given for balancing out the continuous flux.

A rather universal type of bridge for making the necessary impedance measurements in connection with transformer studies is shown. This bridge is adapted to measuring iron core coils, which carry both alternating and continuous currents, when either inductive or capacitive reactive.

Introductory

The present general tendency in the design of the various parts and stages of a radio broadcast receiver is toward the ideal for each part and stage separately and not for the entire set as a unit. For example, the design of the loudspeaker is aimed toward a flat frequency response and a constant impedance without regard to the possibilities of correcting some of the defects in the loudspeaker by means of the amplifier which supplies power to the loudspeaker. Assuming that ideal loudspeakers are available, the aim in designing audio amplifiers for broadcast receivers must also be toward the ideal. It is the purpose of this paper to present material for designing transformers for audio amplifiers of the transformer coupled type which are preassigned characteristics that are considered to be ideal for broadcast reception. The material may also be applied to the design of transformers for the audio amplifier of a continuous wave receiver and for other purposes. The discussion of the impedance and voltage amplification characteristics which is given

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in the next section is intended for the case in which the tube source is an amplifier and not a detector. The amplification characteristics for the case in which the tube source is a detector may be such as to warrant a transformer which is designed especially for the detector stage.

An ideal transformer is defined as one which adapts the impedance of the load to the impedance of the source of power in such a way that the power delivered to the load is a maximum. When the load is a pure resistance an ideal transformer will give rise to a phase shift and an attenuation which are independent of the frequency. For any general load impedance the range of frequencies over which there is substantially no change in phase shift and attenuation is very limited. The general requirements of an ideal transformer are as follows: (1) It must have unity coupling between its windings, or, in transformer language, its leakage inductances must be zero. (2) It must have a ratio of turns which will make the load impedance when transferred to the primary substantially equal to the impedance of the source of power. (3) The impedance of the secondary must be several times the impedance of the load. (4) The effective resistance of the primary and of the secondary windings must be very much smaller than the resistances of the source and load respectively. (5) The windings must be free from the effects of distributed capacitances.

The problems of designing transformers for interstage coupling units in audio amplifiers arise from the fact that it is not feasible to fulfill the above requirements for an ideal transformer. The nature of the load and of the source and the frequency range which the amplifier must cover are such that it is difficult to construct a transformer which will fulfill all of the above requirements. It is impossible to fulfill the first four requirements without getting into serious distributed capacitance effects in the windings. The distributed capacitances of the windings have very little effect upon the behavior of the transformer at the low frequencies, but have a very serious effect at the higher frequencies. These considerations and the fact that phase shift in an audio amplifier for broadcast reception causes no loss in intelligibility, place the design of transformers for audio amplifiers on a basis which is somewhat different from that of the ideal. The aim is toward a reasonable voltage amplification for each stage, i.e., one tube and one transformer, of the amplifier which is independent of the frequency over a range which is necessary for broadcast reception.

THEORY AND DISCUSSION OF IMPEDANCE AND VOLTAGE AMPLIFICATION CHARACTERISTICS

In Fig. 1 is shown an equivalent circuit diagram of a typical transformer for audio amplifiers, with a tube source of power and tube load. This diagram applies without modification only to certain types of transformers at the higher frequencies at which the load current, due to the capacitance load, is several times the exciting current. In this diagram the symbols which are shown have the following definition:

$\mu_1 E_{1G}$ is the voltage generated in the tube source.

R_T is the internal resistance of the tube source.

C_P is the distributed capacitance of the primary winding.

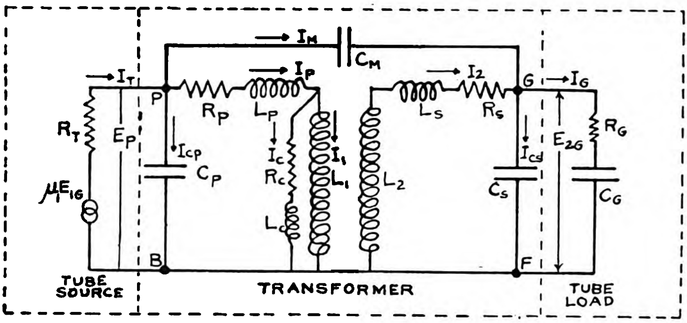

Fig. 1—Equivalent A.C. Circuit of A Transformer with A Tube Source of Power and Tube Load.

R_P and L_P are the resistance and the leakage inductance of the primary.

R_C and L_C are a resistance and an inductance which take care of the core loss and the magnetizing current of the transformer. (See later discussion.) L_1 and L_2 are fictitious inductances which transfer the load current and load voltage of the transformer. The reactances of L_1 and L_2 are always very large compared to all other reactances connected to them. The coupling between L_1 and L_2 is unity. The quantity $\sqrt{I_2/I_1}$ is defined as the ratio of transformation N. N is substantially equal to the secondary turns divided by the primary turns. The numerical value of N may be either positive or negative. N is positive if the directions of the windings are such that currents in the arrow directions, as shown on the diagram, result in fields which aid each other, and negative if the fields oppose each other.

R_S and L_S are the resistance and the leakage inductance of the secondary.

C_S is the distributed capacitance of the secondary.

C_M is the mutual capacitance between the primary and secondary.

R_G and C_G are the equivalent series resistance and capacitance of the tube load.

E_{2G} is the voltage impressed on the grid of the tube load.

The leakage inductances L_P and L_S are due to the load currents which set up fluxes about L_1 and L_2 that link with one coil and not the other. The distributed capacitances C_P and C_S are due to charging currents which flow from layer to layer in the windings of the primary and the secondary coils. C_M is due to the charging current which flows from the outer layer of one winding to the outer layer of the other. C_M may not be present in a particular design. It will be shown in a later section how the leakage inductances and the distributed capacitances can be calculated for a particular design.

At the low frequencies the diagram of Fig. 1 will have to be modified. Usually the sizes of the capacitances C_M, C_S, C_G, and C_P are such that they can be ignored entirely at frequencies below about one third of the first resonant frequency which will be defined later. Also the leakage reactances which are due mainly to the load currents are negligible at the low frequencies below about one third of the first resonant frequency. Then at the low frequencies the transformer acts as an impedance of value $R_P+R_C+j\omega L_C$ in the plate circuit of the tube source. The voltage across the primary is passed on to the secondary in the ratio $E_{2G}/E_P=N$. Consequently it can easily be shown that the ratio of E_{2G} to E_{1G} is given by the expression,

$$\frac{E_{2G}}{E_{1G}}=\mu_1 N\,\frac{\sqrt{R_C{}^2+\omega^2 L_C{}^2}}{\sqrt{(R_T+R_P+R_C)^2+\omega^2 L_C{}^2}} \tag{1}$$

At low frequencies and low flux densities R_C is generally so small compared to R_P that it may be neglected in the denominator of the above expression.

Equation (1) gives definite information on the design for flat voltage amplification at the lowest frequencies. For example, if the voltage amplification of a tube and a transformer is to be equal to or more than 95 per cent of μ_1 N at the lowest frequency, the reactance ωL_C must be at least three times $R_T+R_P+R_C$. This might be stated as one requirement of a good transformer for audio amplifiers, i.e., the reactance of the primary winding must be at least three times the resistance of the tube source of power plus the effective resistance of the primary winding at the lowest frequency for which the amplifier is designed. As an example take the transformer which gave the upper curve of Fig. 4. The inductance of the primary of this transformer is, by measurement, 45 henries. The tube source resistance is 8100 ohms and the resistance of the primary winding is 3250 ohms, which makes $R_T+R_P=11350$. The frequency which is necessary to make $\omega L_C=11350\times 3$ is 125 cycles. This checks very well with the upper curve of Fig. 4.

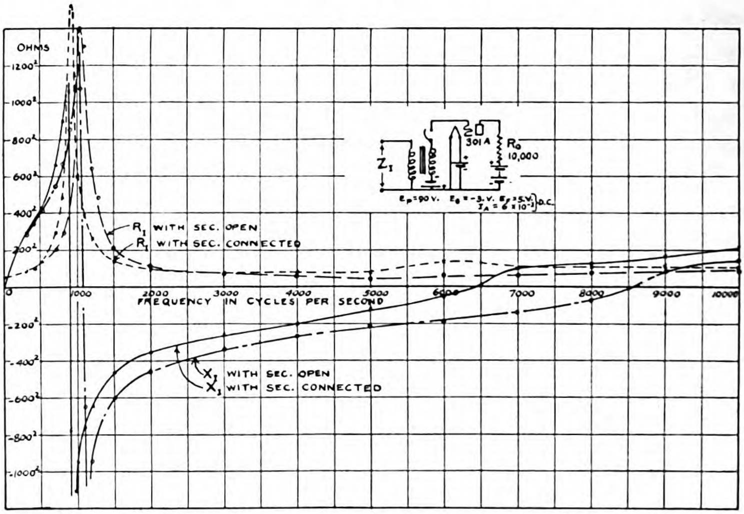

Fig. 2—Typical Impedance vs. Frequency Characteristics of A Transformer with Secondary Open and with Secondary Connected to A Tube as Shown.

At the lowest frequencies the impedance characteristics of the primary of an audio transformer are essentially the characteristics of a resistance and inductance in series because the effective impedance of the load is several times the impedance of the secondary winding. Fig. 2 is a typical set of impedance characteristics of a transformer for audio amplifiers. Consequently the voltage amplification per stage is given almost exactly by (1) as long as the frequency is less than about one third of the first resonant frequency. The upper curve of Fig. 4 shows the voltage amplification characteristics for the same transformer which was used in taking the impedance characteristis of Fig. 2. It will be noted that for frequencies up to about 500 cycles the variation in amplification for this particular transformer is what would be expected from (1). The last tube is terminated in a resistance which is a little higher than the internal resistance of the tube. The amplification of the last tube is constant at 4.7. The input capacitance of the last tube is about constant at 60×10^{-12} farads.

For those transformers which conform very closely to the equivalent circuit of Fig. 1 the impedance of the primary builds up to a very high value around the first resonant frequency, i.e., the first frequency at which the reactance falls through zero. This frequency, which can be derived from later theory, is given to a very close approximation by the expression,

$$f=\frac{1}{2\pi\sqrt{L_C\left[C_P+\frac{(C_S+C_G)C_M}{C_M+C_S+C_G}+(C_M+C_S+C_G)\left(N+\frac{C_M}{C_M+C_S+C_G}\right)^2\right]}} \quad (2)$$

Around this frequency the voltage amplification per stage is substantially equal to $\mu_1 N$ because of the very high impedance which the transformer offers to the tube source. Beyond this frequency and up to the second resonant frequency the combined transformer and load act as a capacitance reactance to the tube source. The impedance falls very rapidly and approaches a value which depends largely upon the resistances of the transformer and tube source.

In order now to explain the impedance and voltage amplification characteristics of the transformer from the first resonant frequency to the highest frequency for which the transformer is useful it will be necessary to develop the equations which apply in this range of frequencies. The two relations which are desired are the expressions for the voltage amplification per stage, E_{2G}/E_{1G}, and the input impedance, Z_1, of the combined transformer and tube load. The expression E_{2G}/E_{1G} is derived in two steps, i.e., E_{2G}/E_P is derived first and then after Z_1 is derived E_P/E_{1G} can be obtained and finally the expression for E_{2G}/E_{1G} is written. These relations are derived under the following assumptions. First, the reactance of the secondary winding, ωL_2 is several times the equivalent reactance of the load. Second, the core losses and magnetizing current are accounted for by

$$I_P = I_1 + I_C = I_1 + \frac{E_P - I_P Z_P}{Z_C} \tag{7}$$

$$I_C = \frac{E_P - I_P Z_P}{Z_C} \tag{8}$$

$$I_R = I_G + I_{CS} = I_Z + I_M = -\frac{E_{2G}}{j\dfrac{1}{\omega C_R}} \tag{9}$$

In the above equations $Z_P = R_P + j\omega L_P$, $Z_S = R_S + j\omega L_S$, $Z_C = R_C + j\omega L_C$ and $C_R = C_S + C_G$.

The above six equations are sufficient to solve for the ratio of E_{2G} to E_P. The solution of these equations yields,

$$\frac{E_{2G}}{E_P} = -\frac{\dfrac{j}{\omega(C_M + C_R)}\left[(-N) + j(Z_S + N^2 Z_P)\omega C_M\right]}{Z_C' Z_S + N^2 Z_P - j\dfrac{Z_C'}{\omega(C_M + C_R)}} \tag{10}$$

In which $Z'_C = 1 + Z_P/Z_C$. Equation (10) gives the voltage transformation of the transformer alone either with the load C_G, or without the load when $C_G = 0$ and $C_R = C_S$. Equation (10) is particularly useful for predetermining the voltage amplification per stage when certain quantities in the expression for the impedance Z_I, which will be pointed out later, are not negligible.

The input impedance Z_I of the transformer with load is derived from (4) to (9) and the additional equation,

$$I_P' = I_P + I_M \tag{11}$$

The solution of these equations gives a value for the admittance Y_I' as follows:

$$Y_I' = \frac{I_P'}{E_P} =$$

$$\frac{Z_C'\left(\dfrac{N}{Z_C'} + \dfrac{C_M}{C_M + C_R}\right)^2}{N^2 Z_P + Z_C' Z_S - j\dfrac{Z_C'}{\omega(C_M + C_R)}} + \frac{1}{Z_P + Z_C} + j\frac{\omega C_R C_M}{C_M + C_R} \tag{12}$$

And finally the value of $Y_I = 1/Z_I$, which is equal to $Y_I' + j\omega C_P$ is given by the expression,

$$Y_I = \frac{Z_C'\left(\dfrac{N}{Z_C'} + \dfrac{C_M}{C_M+C_R}\right)^2}{N^2 Z_P + Z_C' Z_S - j\dfrac{Z_C'}{\omega(C_M+C_R)}} + \frac{1}{Z_P+Z_C} + j\omega\left(\frac{C_R C_M}{C_M+C_R} + C_P\right) \qquad (13)$$

from which the impedance Z_I can be determined. Fig. 2 shows the impedance characteristics of a typical 2 to 1 transformer which has the constants given in Table I. It will be noted that the general shape of the resistance and reactance curves around the second resonant frequency is accounted for mainly by the reciprocal of the first term in the expression for the admittance.

The next step in the procedure is to obtain the ratio of $\mu_1 E_{1G}$ to E_P. This is done by adding the expression for Z_I to R_T to obtain the impedance which acts against the voltage $\mu_1 E_{1G}$. Then

$$E_P = \frac{\mu_1 E_{1G}}{R_T + Z_I} Z_I \qquad (14)$$

or

$$\frac{\mu_1 E_{1G}}{E_P} = \frac{R_T + Z_I}{Z_I} \qquad (15)$$

And finally, since the ratio of E_{2G} to E_{1G} is the expression which is desired it can be written from (10) and (15) as follows:

$$\frac{E_{2G}}{E_{1G}} = \mu_1 \frac{\dfrac{j}{-\omega(C_M+C_R)}\left[(-N) + j(Z_C' Z_S + N^2 Z_P)\omega C_M\right] Z_I}{\left[Z_C' Z_S + N^2 Z_P - j\dfrac{Z_C'}{\omega(C_M+C_R)}\right]\left[R_T + Z_I\right]} \qquad (16)$$

Equation (16) is of such form that it is difficult to gain from it much of an idea how the voltage amplification of the combined transformer and tube source vary with the frequency. It is also too unwieldy to apply to the design of transformers unless long and tedious calculations are made. However, in order to gain an approximate idea of the variation of voltage amplification with frequency it is fortunate that certain quantities in the equation are small compared to others for certain types of design and may therefore be neglected. Suppose, for example, one is interested in the frequency at which the voltage amplification is a maximum and beyond which it drops off very rapidly. An examination of (16) and (13) will suggest that this frequency is near the frequency which will make,

$$N^2Z_P+Z_C'Z_S-j\frac{Z_C'}{\omega(C_M+C_R)}$$

a minimum. Near this frequency the last two terms of (13) can generally be neglected for a first approximation. Neglecting these terms will greatly simplify (16). It will become

$$\frac{E_{2G}}{E_{1G}}=$$

$$\mu_1\frac{j\dfrac{1}{\omega(C_M+C_R)}\left[-N+j(Z_C'Z_S+N^2Z_P)\omega C_M\right]}{R_TZ_C'\left(\dfrac{N}{Z_C'}+\dfrac{C_M}{C_M+C_R}\right)^2+Z_C'Z_S+N^2Z_P-j\dfrac{Z_C'}{\omega(C_M+C_R)}} \qquad (17)$$

The next approximation depends largely upon the kind of magnetic material which is used for the core of the transformer. It will be remembered that Z_C' is equal to $1+Z_P/Z_C$, in which Z_C takes care of the core loss and magnetizing current. If Z_C is at least ten times as high as Z_P the error in calling $Z_C'=1$ will be mainly one of using values for the quantities which are multiplied by Z_C' that are 10 per cent less than they should be. This is not a serious error for the preliminary design of a transformer. The exact quantities can always be calculated after the preliminary design is decided upon. There are also dielectric losses in the distributed capacitances which will tend to make the actual amplification less than the calculated value because these losses will give rise to effective resistances of the windings which are higher than the d.c. resistances.

For the preliminary design and for the purpose of gaining an approximate insight into the part that the windings constants play in the behavior of the transformer Z_C' is set equal to unity. The frequency which makes

$$\omega(L_S+N^2L_P)-\frac{1}{\omega(C_M+C_R)}=0 \qquad (18)$$

is called the second resonant frequency f_r. The absolute value of (17) is then written in the form

$$\frac{E_{2G}}{E_{1G}} = \mu_1 \frac{\frac{1}{\omega(C_M+C_R}\sqrt{\left(N+\frac{C_M}{C_M+C_R}\frac{f^2}{f_r^2}\right)^2+(R_S+N^2R_P)^2\omega^2C^2_M}}{\sqrt{\left[R_T\left(N+\frac{C_M}{C_M+C_R}\right)^2+R_S+N^2R_P\right]^2+\left[X_S+N^2X_P-\frac{1}{\omega(C_M+C_R)}\right]^2}} \quad (19)$$

in which

$$f_r = \frac{1}{2\pi\sqrt{(L_S+N^2L_P)(C_M+C_R)}} \quad (20)$$

For winding schemes similar to Fig. 5 the mutual capacitance C_M will generally be one-fourth, or less, of C_M+C_R. For these schemes then the quantity

$$\frac{f^2}{f_r^2}\,\frac{C_M}{C_MC_R}$$

is small compared to N for all frequencies below f_r and does not play much part in the characteristics of the transformer. Also the quantity $(R_S+N^2R_P)\omega C_M$ will be small compared to

$$\left(N+\frac{f^2}{f_r^2}\,\frac{C_M}{C_M+C_R}\right)$$

and may therefore be neglected. Consequently for winding designs similar to Fig. 5 the following conclusions can be drawn from (19). It makes very little difference whether the coupling between the primary and secondary is positive or negative. For designs in which

$$R_T\left(N+\frac{C_M}{C_M+C_R}\right)^2 + N^2R_P + R_S$$

is less than

$$\frac{1}{\omega_r(C_M+C_R)}$$

the maximum voltage amplification of the transformer and tube source will occur substantially at f_r or the frequency which will satisfy (18). At the frequency f_r the amplification will be substantially equal to

$$\frac{\mu_1 N\frac{1}{\omega_r(C_M+C_R)}}{R_T\left(N+\frac{C_M}{C_M+C_R}\right)^2+R_S+N^2R_P} \quad (21)$$

which may be several times $\mu_1 N$ unless the transformer is carefully designed. Above f_r the amplification will drop off very rapidly. The transformer will be useful as a coupling unit up to the frequency f_r.

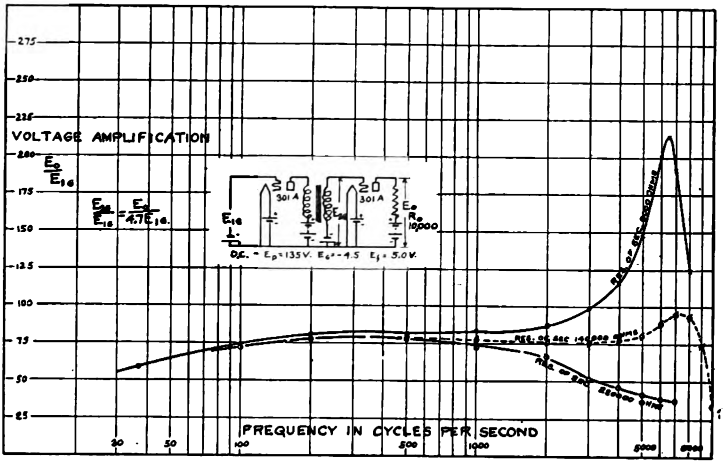

Fig. 4—Amplification vs. Frequency Characteristics Which Show the Effect of Winding Resistance Wire into the Secondary of A Transformer.

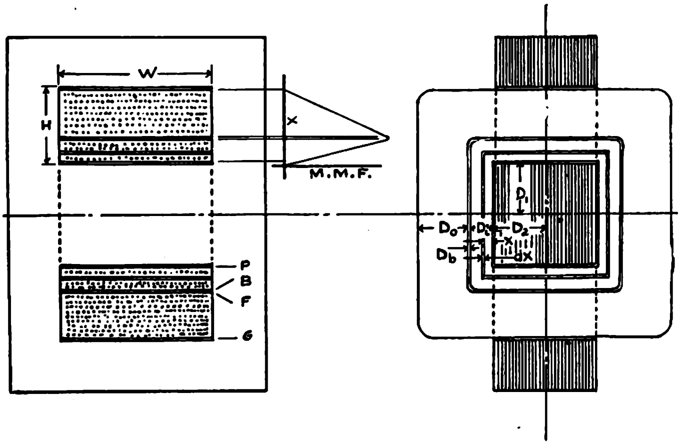

Fig. 5—Simple Winding Scheme for A Transformer.

Fig. 4 shows the amplification characteristics which result from 2 to 1 ratio transformers, with silicon steel cores, which are wound as shown in Fig. 5. For the upper curve the transformer and tubes have the following constants.

TABLE I

$C_M = 28 \times 10^{-12}$ farads, measured value.
$C_S = 60 \times 10^{-12}$ farads, calculated and checked value.
$C_G = 60 \times 10^{-12}$ farads, measured value.
$R_P = 3200$ ohms, measured d.c. value.
$R_S = 8000$ ohms, measured d.c. value.
$R_G = 8600$ ohms which reduced to the effective value in series with R_S becomes 3500 ohms.
$R_T = 8100$ ohms. $\mu_1 = 8.6$
$N_P = 8000$ turns and $N_S = 16000$ turns. $N = +2$.
$N^2L_P + L_S = 3.76$ henries, calculated value.
$N^2L_P + L_S = 3.64$ henries, measured at 1000 cycles.
Area of core equals 5.75 square centimeters.
Mean length of path equals 14.4 centimeters.

Dimensions of winding space are W equals 3.6 cm and H equals 1.2 cm. The core losses will increase the effective resistance of the transformer less than 10 per cent of the combined d.c. resistances of the primary and the secondary when referred to the secondary. Using the measured value of leakage inductance and the above values for the capacitances f_r calculates to be 6800 cycles. The sum of the capacitances, $C_M + C_S + C_G$, was checked experimentally. Upon substituting the above constants into (19) there results $E_{2G}/E_{1G} = 47.2$ at 6600 cycles under the assumption that the true value of f_r is 6600 cycles. From the upper curve of Fig. 4 the overall voltage amplification at 6600 cycles is 213. The amplification of the last tube is 4.7. Consequently $E_{2G}/E_{1G} = 213/4.7 = 45.4$ which is to be compared with 47.2. This check serves to show that the assumptions and approximations are well within reason for the preliminary designs similar to the transformer which is used for the above data.

There are some interesting things revealed by (19) and (1). Suppose, for example, it is desired to redesign the transformer which is used for the above data so that the voltage amplification is substantially equal to $\mu_1 N$ up to f_r. One method which the equations and the equivalent circuit suggest is to increase the resistance of the secondary winding by replacing the copper wire with resistance wire. This will not affect the amplification at the low frequencies but will have a desirable effect around the frequency f_r. It is also expected that winding a part of the coil with resistance wire and the rest with copper wire will have exactly the same effect as though the same total amount of resistance were distributed throughout the coil. Calculations show that it will require about 140,000 ohms to make the characteristics flat. The appropriate curve of Fig. 4 shows the voltage amplification characteristics which result from a transformer that is wound with resistance wire in the secondary. The inner three quarters, by turns, of the secondary coil are wound with copper wire and the outer one-quarter with resistance wire having 60 times the resistance of copper. The total resistance of the secondary coil is 140,000 ohms. The other constants of this transformer are not quite the same as those of the original design because the peak is shifted from 6500 to 7500 cycles. The results which are shown serve to show the advantages of using the resistance wire on the secondary coil. The lower curve in the same figure shows the effect of increasing the resistance to 280,000 ohms.

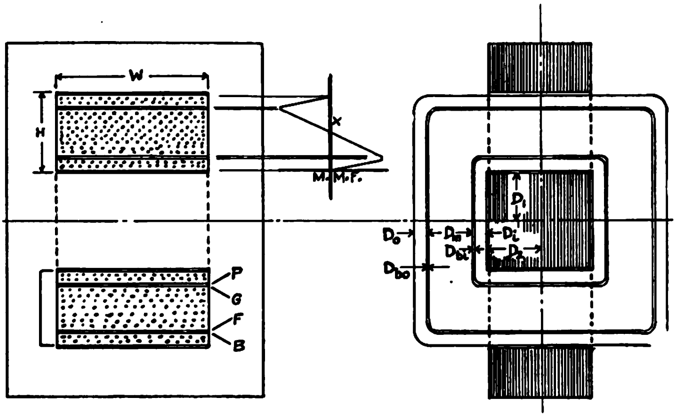

Fig. 6—Winding Scheme for Low Leakage Inductance.

Another thing which (19) suggests is to wind the transformer in such a way that it will have a low leakage inductance. If the winding scheme is of the type shown in Fig. 6 the reduction in leakage inductance is attended by an increase in the mutual, or effective secondary, capacitance no matter how the leads are brought out from the coils. This, however, is not an undesirable feature because with a relatively large effective capacitance across the secondary the action of the transformer will be rather independent of the tube load. After examining the various schemes including the possibilities of splitting either the secondary or the primary winding, the scheme which is shown in Fig. 6 was selected. This scheme gives rise to a large mutual capacitance C_M, but by placing the plate and grid layers next to the inner space between the coils instead of the outer space and by increasing this space the capacitance can be kept comparatively small. For transformers in which the mutual capacitance is the major por-

only 235×10^{-12} farads. The characteristics of curve A are the best for radio broadcast reception because the transformer cuts off very rapidly after 5000 cycles.

CALCULATION OF THE TRANSFORMER CONSTANTS

In the theory on the impedance and the voltage amplification characteristics of transformers several constants are used. For a given transformer most of these constants can be measured but when it comes to laying out a new design one must be able to predetermine the constants from which the performance of the transformer can be predicted approximately. It is the purpose of this section to present relations from which the design constants of the transformer can be calculated. Some of the relations are taken from various sources, but are presented for the sake of completeness.

In transformer theory it is customary to take care of the core losses and the magnetizing current by a resistance R and an inductance L in parallel across the induced voltage of the transformer. This is justified because the core losses are nearly proportional to the square of the induced voltage and the magnetizing current is proportional to the induced voltage. The inductance L can be calculated by the expression,

$$L = \frac{4\pi 10^{-9} N_P{}^2 \mu_r A}{l} \tag{22}$$

in which A is the net area of the core and l is the mean length of the magnetic path, if the relative permeability μ_r of the core material is known at the proper alternating and continuous flux densities. Curves and information similar to that which is given in Fig. 10 are useful for determining the value of L.

The resistance R can be calculated if the hysteresis and eddy current coefficients, K_h and K_e, of the material are known. Since the core losses in audio-frequency transformers are of minor importance and since it simplifies matters the total core losses are assumed to be proportional to the square of the a.c. flux density and are given by,

$$P = V(K_h f + K_e f^2) B^2$$

This is correct for the eddy current losses but only approximate for the hysteresis losses. Then if a new constant K_c is defined by the relation,

$$K_C = K_h f + K_e f^2 = \frac{\text{Total core loss per cu. cm.}}{B^2} \qquad (23)$$

it can be shown that

$$R = \frac{E_i^2}{P} = \frac{2\pi^2 10^{-16} f^2 N_P^2 A}{K_C l} \qquad (24)$$

The constant K_c can be determined experimentally from test samples at representative values of alternating flux densities and continuous magnetizing forces. If the hysteresis and eddy current losses are separated the constants K_h and K_e can be determined and K_c can be calculated for any frequency.

The hysteresis constant K_c depends entirely upon the kind of material and is independent of any dimensions. For 3.5 per cent silicon steel, which has been used to a large extent for core material, it has the approximate value of 80×10^{-12} when B is in maxwells per square centimeter. At the time of this writing very little has been published on the hysteresis losses in the newer core materials of high permeability at low flux densities. The eddy current constant K_e for laminated cores such as are used in audio-frequency transformers depends upon the conductivity of the material and the thickness of the sheet. When B is in maxwells per square centimeter K_e is given by the equation,

$$K_e = \frac{\pi^2 \gamma t^2}{6 \times 10^{16}}$$

in which γ is the conductivity and t is the thickness of sheet. For 3.5 per cent silicon steel $\gamma = 2 \times 10^4$ mhos per centimeter cube. For "*A*" metal $\gamma = 2.2 \times 10^4$ mhos per centimeter cube.

The resistance R_c and the inductance L_c are then given by the expression,

$$R_C = \frac{R}{1 + \frac{R^2}{\omega^2 L^2}} \quad \text{and} \quad L_C = \frac{L}{1 + \frac{\omega^2 L^2}{R^2}} \qquad (25)$$

At low frequencies and low a.c. flux densities the equivalent core loss resistance R, for most transformers which are suitable for interstage coupling units, is large compared to ωL and L_c is substantially equal to L.

The calculations of the d.c. resistances R_p and R_s of the windings are well known and need no further discussion. The leakage inductances L_p and L_s are determined from expressions which are taken from electrical texts and which are found to give very good results under certain conditions. Referring to Fig. 5 the leakage inductance expression for the type of winding which is shown is based on the assumption that the flux in a shell dx thick and having the dimensions shown is distributed uniformly over the area and is due to the load ampere turns inside the shell. It is also assumed that the reluctance is equal to W divided by $2\pi 10^{-9}$ times the area of the shell and that the reluctance of the return path through the iron is negligible. It is also assumed that the distribution of the magnetomotive force over the end of the coil is as shown in the figure. Under these assumptions the leakage inductance in henries, when all dimensions are in centimeters, for the winding in Fig. 5 is, for the primary,

$$L_P=\frac{16\pi N_P^2}{10^9\omega}\left[\left(\frac{D_1+D_2}{3}+\frac{D_i}{2}\right)D_i+\frac{1}{2}(D_1+D_2+2D_i+D_b)D_b\right]$$

and for the secondary,

$$L_S=\frac{16\pi N_S^2}{10^9\omega}\left[\left(\frac{D_1+D_2+2(D_i+D_b+D_0)}{3}-\frac{D_0}{2}\right)D_0\right. \tag{26}$$

$$\left.+\frac{1}{2}(D_1+D_2+2D_i+D_b)\,D_b\right]$$

For a winding of the type which is shown in Fig. 6 the leakage inductances are,

$$L_P=\frac{4\pi N_P^2}{10^9\omega}\left[\left(\frac{D_1+D_2}{3}+\frac{D_i}{2}\right)D_i\right.$$

$$\left.+\left(\frac{D_1+D_2+2(D_i+D_{bi}+D_{b0}+D_m+D_0)}{3}-\frac{D_0}{2}\right)D_0+\frac{S}{2}\right]$$

$$L_S=\frac{4\pi N_S^2}{10^9\omega}\left[\left(\frac{D_1+D_2+2D_i+2D_{bi}+D_m}{3}\right)D_m+\frac{S}{2}\right] \tag{27}$$

In both of the above expressions,

$$S=(D_1+D_2+2D_i+D_{bi})D_{bi}+(D_i+D_2+2D_i+2D_{bi}+2D_m+D_{b0})D_{b0}.$$

For most transformers the quantities which are multiplied by D_b, D_{bi}, and D_{bo}, which are due to the leakage flux between the windings, are negligible. There are cases, though, for a winding of the type which is shown in Fig. 6 in which one of these spaces is deliberately increased to reduce the mutual capacitance between windings. For these cases an appreciable error might be made by neglecting the leakage flux in this space.

One thing in particular to be noted is the large reduction in leakage inductance which is effected by employing a winding of the type shown in Fig. 6 instead of the usual type which is shown in Fig. 5. This reduction in leakage inductances is attended by an increase in the mutual capacitance as suggested by the manner in which the leads are brought out. The mutual capacitance caused by a winding of this type will be the major portion of the effective capacitance across the secondary and will be decreased directly as D_{bi} is increased. The leakage inductance due to the flux in the space D_{bi} will be a minor portion of the total leakage inductance as long as D_{bi} is only one-tenth as large as D_i or D_o. Consequently a winding of this type has merits over the simple type which is shown in Fig. 5 because the product of the effective capacitance and total leakage inductance referred to the secondary, which determines approximately the second resonant frequency, can be kept smaller for this type than for the usual type of winding of the same total turns and ratio.

Another type of winding which might be used and which will be discussed in the next section is shown in Fig. 9. It is impossible to derive an expression for the leakage inductance for a transformer wound according to this type that is anything more than a rough approximation. An expression which has been found to give approximate results is based upon the following assumptions. The flux in a pancake section dx units thick is due to the load ampere turns to the right, or the left, of the section and is distributed uniformly over a mean area of $dx.\ MT$ square centimeters, in which MT is the mean length of turn in the coil in centimeters. The reluctance of the path is equal to the distance H divided by $4\pi 10^{-9}$ times the mean area. These notions give the following expressions for the leakage inductance in henries, when all dimensions are in centimeters, of the split secondary type of winding.

$$L_P=\frac{\pi N_P^2}{10^9 H}MT.\left[\frac{D_m}{3}+\frac{D_{b1}}{2}+\frac{D_{b2}}{2}\right]$$

and

$$L_S=\frac{\pi N_s^2}{10^9 H}MT.\left[\frac{D_r+D_l}{3}+\frac{D_{b1}}{2}+\frac{D_{b2}}{2}\right] \qquad (28)$$

The distributed capacitances of the windings for any type of coil are composed of the layer to layer capacitances and the turn to turn capacitances. For multi-layer coils in which the number of turns is many times the number of layers the turn to turn capacitances can generally be neglected. The equivalent capacitance which must replace all of the series of layer to layer capacitances can be arrived at from the charging current which flows from layer to layer and does not contribute to the magnetic field about the coil. Suppose the winding of all layers is started from the same end of the coil. Then, if each layer consumes E volts, there would be E volts across each layer to layer condenser. If J is the number of layers, there would be $J-1$ such condensers in series. These $J-1$ condensers would be acting under $(J-1)$ E volts. The total voltage across the coil would be JE volts. So that the series capacitance which is equal to $C_l(1/J-1)$ would have to be reduced by the factor $(J-1)/J$ to get the equivalent condenser across the terminals of the coil. Finally then, the approximate capacitance across the terminals is $C_t = C_l/J$ in which C_l is the capacitance between the mean layers. This is substantially equivalent to the actual winding case in which there are $2E$ volts between two successive layers at one end of the coil and zero at the other end because the average charging current which flows through the winding from layer to layer is substantially the same for the actual arrangement and the arrangement pictured above. The mean layer to layer capacitance for a winding of the type shown in Fig. 5 is given by the expression,

$$C_l = p\frac{MT.(\omega - u)}{D_p + D_e} \tag{29}$$

in which p equals 22×10^{-14} for paraffin paper, $W-U$ is the actual length of the winding in centimeters, MT is the mean length of turn in centimeters, D_p is the thickness of the paper, or other insulation, between the layers in centimeters, and D_e is twice the thickness of the insulation on the wire in centimeters.

The calculation of the mutual capacitance C_M depends entirely upon the type of winding. If the transformer is wound according to Fig. 5 the mutual capacitance is the capacitance between the outer layer of the secondary and the inner layer of the primary coils. A metal case placed around the transformer and connected to the F terminal will eliminate practically all of the mutual capacitance but will increase both the primary and secondary capacitances by small amounts which can be estimated. For a winding of the type which is shown in Fig. 6 the mutual capacitance can be calculated very closely by (29) except that MT is replaced by the proper dimension and p is assigned a value which depends upon the kind of insulation. For this type of winding the mutual capacitance will generally be the major portion of the effective capacitance across the secondary.

DISCUSSION OF THE DESIGN RELATIONS FOR TRANSFORMERS

In the previous sections of this paper are given the necessary relations from which a transformer can be designed and its performance predicted when it is used with a given tube load and tube source. The procedure in designing a transformer for an amplifier with certain preassigned characteristics is one of successive trials by calculation. Due regard must be given to the principles of production. Usually a designer has his own past experience or the experiences of others as a starting point and guide in laying out a new design. The procedure in a general way is to fix the core area, the number of turns, and the length of the magnetic path for a given performance at the lowest frequency by applying (22) for the inductance and (1) for the performance after the resistance of the primary has been calculated. Then a type of winding is selected and the leakage inductance, the secondary resistance, and the capacitances are determined. Finally the approximate performance at the high frequencies is calculated from (19), (20), and (21). After this preliminary design is completed and studied it will become apparent what changes are necessary in the transformer in order to make each stage of the amplifier perform according to the preassigned characteristics. When the design conforms to the preassigned characteristics at the higher frequencies according to the approximate relations the more exact performance may be calculated from the relations which take into consideration the distributed capacitance of the primary winding, the core losses, and the magnetizing current. However, there are some general ideas which are guides in the above procedure. Some of these ideas will be discussed briefly.

In considering the shape of the space into which the wire is to be wound about the core one is first led to a square shape because this shape has the minimum length of path. However, there are other considerations which are favorable to a rectangular shape in which, for windings of the type shown in Figs. 5, 6, and 8, W is greater than H. Suppose that as W is increased and H decreased the number of turns and core area are held constant. The first thing to be noted is a decrease in the resistance of the winding. This is desirable for the primary winding. The second thing to be noted is that both the alternating and continuous ampere turns per centimeter are decreased because the length of the path is increased. The former tends to decrease L_c while the latter tends to increase L_c by increasing μ_r. The result is that for most transformers L_c will be decreased less than 15 per cent in changing from a square winding space to one in which W is three times H. This percentage may even be less for a high permeability core. Consider as an example a transformer which has "A" metal for the core. Referring to the lower curve of Fig. 10, suppose for example the continuous ampere turns per centimeter are 1 or less for a square winding space. The winding space is now made rectangular so that the length of the path is increased as much as 50 per cent. Due to approximately a 30 per cent reduction in the continuous ampere turns per centimeter the a.c. permeability is increased almost 50 per cent, so that the primary winding has the same inductance as before the changes were made. This is, of course, an extreme case which is used merely to illustrate the point. It would not be equally true for a silicon steel core nor for the same high permeability material when operating between 1 and 2 continuous ampere turns per centimeter. It is appreciated, of course, that the core losses will be higher for the rectangular winding space than for the square

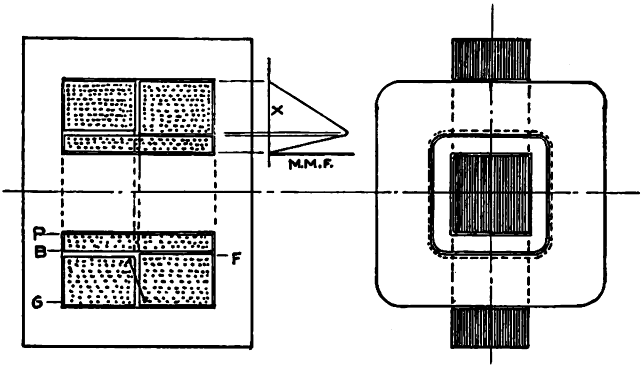

Fig. 8—Winding Scheme for Low Effective Capacitance Across Secondary.

space because it will require more iron in the magnetic circuit. Usually at low frequencies the core losses are small compared to the copper losses and have little effect upon the performance of the transformer. At the high frequencies it might even be desirable to increase the core losses to help limit the amplification at the resonant frequency.

The performance of the transformer at frequencies near the second resonant frequency is also influenced by the shape of the winding space. For coils of the type which are shown in Figs. 5, 6, and 8, as W is increased and H decreased such that the area of the winding space remains constant the distributed and mutual capacitances are increased. The mutual capacitance increases approximately as the first power of W. The effective distributed capacitance increases at a rate slightly greater than the first power of W. The total leakage inductance of the primary and the secondary referred to the secondary decreases faster than $1/W^2$. Consequently the product of the total leakage inductance and the effective secondary capacitance becomes smaller, as W is increased and H decreased, in a favorable manner; i.e., the leakage inductance decreases and the effective capacitance increases. This means that the second resonant frequency increases and that the voltage amplification per stage near the second resonant frequency decreases. For high ratio transformers this effect of making the winding space rectangular is more important than the other two effects which are pointed out above. Therefore, in conclusion it may be stated as a guide in starting a new design that for designs similar to those which are pictured in Figs. 5, 6, and 8, W should be at least three times H.

In Fig. 9 is shown a winding scheme which is especially designed for low effective secondary capacitance and low leakage inductance. This scheme differs from the other ones which are shown in that as dimension H is increased and W decreased the leakage inductance and effective secondary capacitance are both decreased. One bad feature of the scheme is the comparatively high resistance of the primary for a transformer of this type which has the same mean length of path, core area, and number of turns as one of the other designs. Another objection to this type is from a production standpoint. It is more difficult to wind coils of this type than of the other types. However, it is believed that this scheme of winding has merits for a high ratio transformer with high permeability material for the core because the winding space can be shaped for a very low product of leakage inductance and effective secondary capacitance.

Consider now the shape and relative dimensions of the core. Referring to the diagram in Fig. 5, $2D_1$ is the width of the lamination and $2D_2$ is the width of the stack. The first things to be noted are that the leakage inductances, the effective secondary capacitance, and the resistances of the windings are smallest for a square core section. When there is no continuous magnetomotive force present increasing the width of the stack and decreasing the width of the lamination will result in an increase in the self inductance of the primary because of the decrease in the length of the magnetic path. With continuous magnetomotive force present the situation is different and the reduction in the length of the path might easily be offset by a reduction in the a.c. permeability if the width of stack is increased and the width of lamination decreased as the core area is held constant. This depends largely upon the kind of core material. In conclusion, then, for a starting point at least a square section should be adopted. For most cases this shape will be as good as any other shape from a performance standpoint.

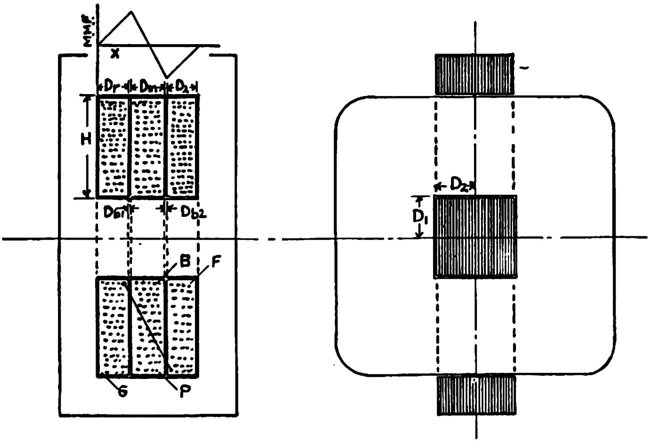

Fig. 9—Winding Scheme for Low Distributed Capacitances and Low Leakage Inductances.

The practice of winding transformers with number 40 enamel copper wire and placing insulation between the layers which has a thickness equal to about one-fifth of the diameter of the wire is reasonable. There are designs, especially those which are similar to Fig. 5, which will be better electrically if smaller than number 40 is used on the secondary. The chief objection to using smaller than number 40 is from a production standpoint.

With the advent of high permeability core materials, which are gradually replacing silicon steel, audio amplifier transformers with ratios of 4 or 6 are being built with just as good frequency characteristics as 2 to 1 silicon steel transformers. The worth of a material which has a permeability at least four times that of silicon steel is appreciated, by an example of the following kind. By replacing the silicon steel core of the 2 to 1 transformer which results in curve *A* of Fig. 7 by a material which has four times the permeability this transformer can easily be converted to a 4 to 1 transformer which will give rise to about the same amplification versus frequency characteristics. This is accomplished by removing half of the primary turns. The result is a transformer which has the same effective secondary capacitance, a slightly lower leakage inductance, and the same primary inductance. The continuous ampere turns per centimeter for a CX301A tube will be approximately 1.2. The "*A*" metal which results in the curve of Fig. 10 has about the right permeability. By the scheme which is given in the next section, this transformer can easily be redesigned into an 8 to 1 transformer.

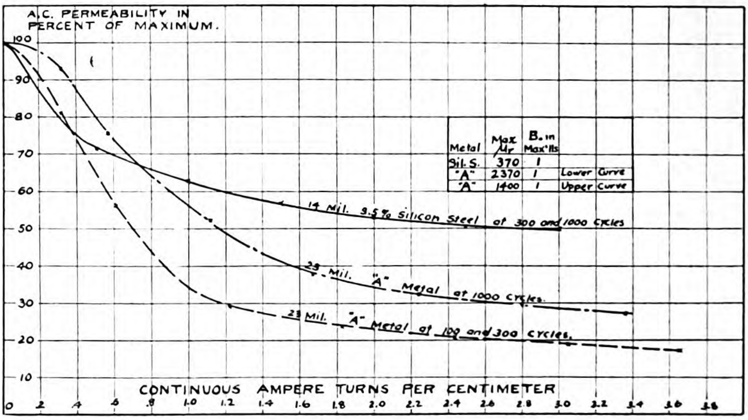

Fig. 10—The Per Cent Change in A.C. Permeability with Continuous Magnetizing Force for Transformer Core Materials at Very Low A.C. Flux Densities.

A METHOD FOR BALANCING OUT THE CONTINUOUS FLUX IN THE CORE OF A TRANSFORMER

The a.c. permeability of the high permeability materials is reduced so much in per cent by the continuous flux in the core that the feasibility of a third winding on the core of the transformer for balancing out the continuous flux suggested itself to the writer. This winding is used for setting up a continuous flux in the core which is equal and opposite to the continuous flux set up be the continuous plate current in the primary winding. In Fig. 11 is shown the desirable effect which this scheme has on a 4 to 1 transformer which has a high permeability core. The winding may be designed to be excited from either the *A* or *B* battery of the receiving set. The winding and circuit are designed so that the resistance R_a of the auxiliary circuit is at least five times the reactance X_a of the auxiliary winding at the lowest frequency. If the permeability were not affected, this will increase the effective resistance of the primary by $K^2X_a^2/R_a$, where K is the ratio of primary turns to auxiliary winding turns. It will decrease the reactance of the primary by 4 per cent or less. The continuous current is regulated so that $I_pN_p = I_aN_a$. As the frequency increases, the reactance X_a increases so that ultimately complete transformer action will take place, and R_a will be transferred as K^2R_a. Consequently K^2R_a must be large compared to the impedance which the combined transformer and load would have, without the auxiliary winding, at the higher frequencies.

The result of balancing out the continuous flux by a properly designed auxiliary winding is well illustrated in Fig. 11. Suppose, however, the application of this principle is made in another way. There are 4000 turns on the primary of this 4 to 1 transformer. Let 1700 of these turns be used for an auxiliary winding and the other 2300 used for the primary, making the ratio about 7 to 1. The per cent change in amplification with frequency will be about the same for the new design, when the continuous flux is balanced out, at the low frequency around 100 cycles as for the original 4 to 1 transformer with continuous flux present. The overall amplification will be 1.75 times as much for the new design as for the original. The second resonant frequency of the new scheme will be slightly higher than that of the original because the effective secondary capacitance is undisturbed whereas the total leakage inductance referred to the secondary is slightly less for the new scheme. The primary resistance and tube source resistance referred to the secondary will be increased because of the higher ratio. This will improve the characteristics at the high frequencies. In order to balance out the continuous flux it will require 23/17 as much continuous current in the auxiliary winding as in the primary. For a CX301A tube source this will be about 7×10^{-3} amperes. In order to obtain this current from the *B* battery or *B* eliminator of 135 volts it will require 19,300 ohms in series. At the high frequency, when X_a is several times R_a, this value of resistance will be transferred into the primary as 35,300 ohms which is about 4 times the internal plate resistance of a CX301A tube. At 100 cycles the reactance of the auxiliary winding is about 4500 ohms when the continuous flux is balanced out of the core. So that the resistance of the auxiliary circuit is about 4.3 times the reactance of the auxiliary coil at 100 cycles. These values are about right for obtaining the desired effect.

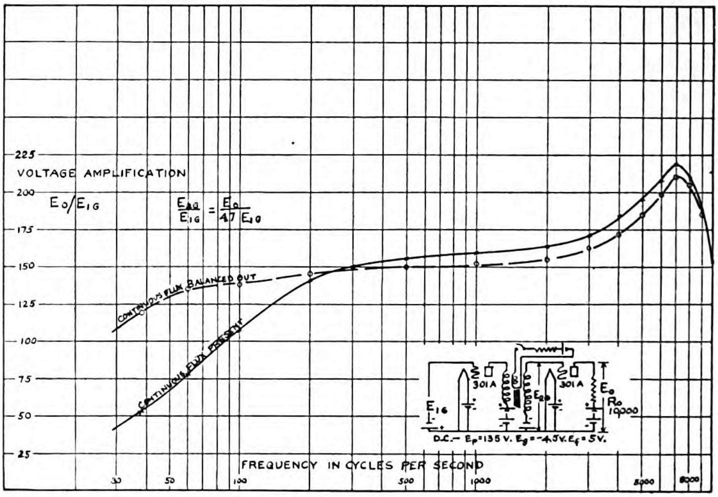

Fig. 11—Amplification vs. Frequency Characteristics Which Show the Effect of Balancing Out the Continuous Flux in the Core of A Transformer.

MEASUREMENT OF THE TRANSFORMER CONSTANTS

In connection with transformer design studies it is essential that the designer have at his disposal some method of measuring the transformer constants. It is appreciated that any method for measuring the impedances of the windings must be of such a nature that the impedance can be measured for a particular value of alternating voltage, or current, and of continuous magnetizing current. The method must also be adapted to measure both capacitive and inductive reactance. The bridge scheme of Fig. 12 has been developed to meet these needs. The scheme has been so worked out that the bridge can be balanced without disturbing the continuous current through the transformer. The continuous current is measured by means of an ammeter in the balance indicator circuit. The alternating voltage across the transformer is measured by a detector voltmeter across the bridge arm which is adjacent to the transformer. This voltmeter should not be left on when the final balance is made.

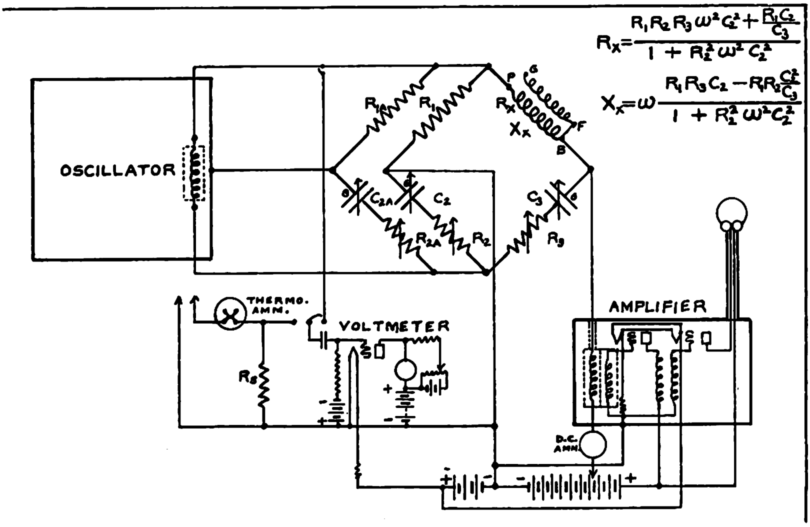
$R_x = \frac{R_1 R_2 R_3 \omega^2 C_2^2 + \frac{R_1 C_2}{C_3}}{1 + R_2^2 \omega^2 C_2^2}$
$X_x = \omega \frac{R_1 R_3 C_2 - R_1 R_2 \frac{C_2^2}{C_3}}{1 + R_2^2 \omega^2 C_2^2}$
OSCILLATOR
R_{1A}
R_1
P
G
R_x
X_x
F
B
C_{2A}
C_2
C_3
R_{2A}
R_2
R_3
THERMO. AMM.
VOLTMETER
R_5
AMPLIFIER
D.C. AMM.
+
-

Fig. 12—Bridge Scheme for Measuring Iron Core Coils.

The purpose of the auxiliary, or guard, arm is to bring the oscillator shield to the same potential as the amplifier shield and detector voltmeter. This is essential when the impedance which is being measured is of the same order of magnitude as the spurious capacitive reactances which exist between the source of power and apparatus connected to the balance points of the bridge. This auxiliary bridge, to be effective, must also be balanced at the same time the main bridge is balanced. Usually it is sufficiently accurate to assume that the auxiliary bridge is balanced when $R_{1A} = R_1$, $C_{2A} = C_2$, and $R_{2A} = R_2$. For greater accuracy, the auxiliary bridge can be balanced by observing with a balance indicator across the balance points. For still greater accuracy the different arms of the bridge can be shielded according to methods which have been given in recent publications provided the constants of the circuit are known after they are shielded. It is believed though that the scheme given is sufficiently accurate for transformer design studies.

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A BRIDGE CIRCUIT FOR MEASURING THE INDUCTANCE OF COILS WHILE PASSING DIRECT CURRENT*

By

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Summary—*A bridge circuit is described in which the inductance of a coil is compared to resistances and a capacitance. A brief comparison is made to other similar circuits.*

THERE has been considerable discussion recently in regard to methods of measuring the inductance of iron core coils while passing direct current. This discussion has brought out a variety of circuits using voltmeter ammeter methods. These give sufficiently accurate results for factory measurements, but where greater accuracy is desired a bridge method is much more desirable. However, the method of comparison to a standard inductance is not very practical in this case. The required high value of inductance is difficult to obtain without an iron core. An iron core inductance is obviously unsuitable for a laboratory standard. Obtaining and measuring the required value of direct current is also a problem that must be met.

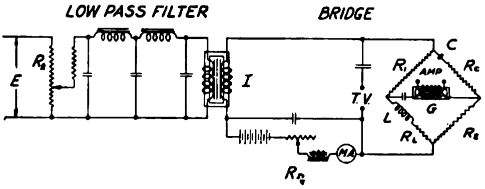

Fig. 1

In the bridge about to be described neither of these problems presents any difficulty. The inductance is measured by comparison with standard resistors and a capacitor. No standard inductor is required. The direct current requires only a direct measurement since its path is not divided.

The circuit of this bridge is given in Fig. 1. It will be seen that the voltage drop across the inductance is balanced against

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* Original Manuscript Received by the Institute, August 24, 1928. Revised Manuscript Received, October 16, 1928.

the drop across a resistance. The phase is corrected by the impedances in the other two legs. Mathematically:

$$\frac{E\,(R_L+j\,X_L)}{R_1+R_L+jX_L} = \frac{E\quad R_2}{R_2+Rc-jXc}$$

Solving for the impedance of the coil,

$$R_L + jX_L = R_1\,R_2\,\frac{Rc + j\,Xc}{Rc^2+Xc^2}$$

$$R_1 = \frac{R_1\,R_2\,Rc}{Rc^2+Xc^2}$$

$$X_L = \frac{R_1\;R_2\;Xc}{Rc^2+Xc^2}$$

$$L = \frac{R_1\;R_2\;C}{1+\left(\frac{Rc}{Xc}\right)^2}$$

It will be noticed that where the resistance of the coil is negligible R_C will be negligible and

$$L = R_1 R_2 C$$

The bridge is then balanced for all frequencies. Where the resistance is not negligible the value R_1R_2C must be divided by the correction factor $\left(1+\left(\frac{Rc}{Xc}\right)^2\right)$. The bridge is then balanced at one frequency only.

In operation R_D is used to adjust the direct current to the correct value and R_A is used to adjust the alternating current to the desired value. It will be realized that these two operations are quite necessary if accurate results are to be obtained. The effective a. c. inductance of iron core coils varies over quite wide limits when the value of the direct current is changed. The value and frequency of the applied voltage also affects the result to a marked degree. In comparing measurements of such coils it is therefore quite necessary to state the value of the direct current and the value and frequency of the alternating voltage.

Any or all of the units R_1, R_2, and C may be varied to obtain an approximate balance. However, it is most convenient to use

R_2. The accurate balance is then obtained by adjusting both R_2 and R_C until zero signal is fed to the amplifier.

It will be found that when R_2 is adjusted alone, a point of minimum signal is found. R_C is then adjusted and a further reduction made in the signal strength. However, the signal is not yet balanced out completely. When R_2 is adjusted while R_C is slightly off from the correct value, the point of minimum signal is slightly different from the point of zero signal as obtained when R_C has the correct value. Hence, it is necessary to adjust first one and then the other back and forth several times to obtain the correct balance.

It is best to use a fairly small value of R_1 for two reasons. First, if R_1 is a good deal smaller than X_L it may be assumed that the a. c. voltage across the coil is the same as that measured across the input to the bridge. Secondly, less battery is required to supply the direct current when R_1 is small. However, if the ratio of R_1 to X_L is made too large the errors due to various capacities to ground may be magnified. The condenser should be a high quality mica standard condenser. The resistance of such a condenser may be neglected for most purposes.

The transformers marked G and I should both be of the rectangular core type, with primary and secondary on opposite legs so as to minimize undesired capacity. When this is done no correction for stray capacities is necessary to obtain very good accuracy at low frequencies.

The low pass filter is a very desirable feature of the circuit, its purpose being to better the wave form of the applied voltage. If the filter is not used the point of balance is disguised by the presence of harmonics which do not balance at the same settings. A skilled ear can easily differentiate between the fundamental and its harmonics and can tell when the fundamental goes out. However, the apparent sharpness is greatly increased if the value of the harmonic content is kept quite low.

The filter should be so adjusted as to pass the fundamental but not its harmonics. Of course it is necessary to change the characteristics of the filter if the frequency is changed. If a source of alternating current is available, the output of which has a very low harmonic content, the filter becomes unnecessary.

The usefulness of this bridge is not confined to filter choke measurements. It is invaluable in measuring the impedance of loudspeaker windings at various frequencies. It should be observed that by merely reversing the positions of R_1 and the unknown, this bridge becomes the conventional one for measuring an impedance having resistance and capacity components. With this in mind the setup may be regarded as a general purpose bridge for measuring impedances of any value or power factor.

While this bridge was developed independently by the writer it has probably been used before for various other purposes. It is of interest to note that a very similar bridge for inductance measurements was originated by Maxwell. This is shown in Fig. 2.

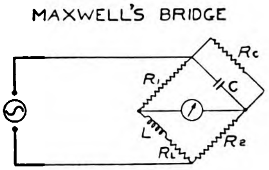
MAXWELL'S BRIDGE

Fig. 2

Theoretically, this bridge has an advantage over that of Fig. 1, in that it balances for all frequencies at the same settings of the controls. The equations for a balance are:

$$L = R_1 R_2 C$$

$$R_L = \frac{R_1 R_2}{R_C}$$

Apparently this bridge is balanced at all frequencies, but unfortunately this does not hold in practice because of the wide variations in the value of R_L when the frequency is changed. Hence, the theoretical advantage of the circuit of Fig. 2 is not a real advantage. In a practical case one circuit is about as far out of balance as the other when the frequency is changed.

This same argument applies to the Anderson bridge shown in Fig. 3. This bridge is a modification of that of Fig. 2 in that the resistor r is added. The equations for balance are:

$$\frac{R_M}{R_N} = \frac{R_X}{R_P} \text{ and}$$

$$L_X = R_M C \left[r\left(1 + \frac{R_P}{R_N}\right) + R_P \right]$$

The reason for the addition of the resistance r is to add a variable which enters into the second equation but not into the first. The idea is that the first equation may be balanced on direct current and then the second may be balanced on alternating current without upsetting the first.

Of course, however, it is found that R_X is of an entirely different order of magnitude on alternating current and on direct current. Hence, it is necessary to readjust the first equation, which upsets the second, and so on.

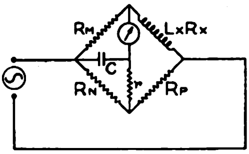
THE ANDERSON BRIDGE

Fig. 3

Thus, with any one of the three circuits shown, in practice it is necessary to make a long series of readjustments of each of two controls to obtain an accurate balance.

It is believed that the Anderson bridge has replaced that of Maxwell to an undeserved extent. The simplicity of the circuits of Fig. 1 and Fig. 2 is a very desirable feature, and the advantages of the Anderson modification very dubious.

The advantage of the circuit of Fig. 1 is the elimination of the divided d. c. path.

Volume 16, Number 12 *December, 1928*

Discussion on
RECENT DEVELOPMENTS IN LOW POWER AND BROADCASTING TRANSMITTERS*

(I. F. Byrnes)

Edward L. Nelson†: In one of the early paragraphs of his paper, Mr. Byrnes discusses the rating of radio transmitters and stresses the importance of excluding losses in the antenna coils in computing the output power. This is a matter which has, at times, occasioned some confusion and his position on this point is well taken. Where radio telephone transmitters are concerned, however, there is another factor of prime significance from a rating standpoint, which is even more frequently neglected, that is, modulation capability. The modulation capability of a transmitter may be defined as the maximum degree of modulation possible without serious distortion, employing a single-frequency sine-wave input and using a rectifier coupled to the antenna in conjunction with an oscillograph or harmonic analyzer to indicate the character of the output. Present day transmitters show wide variations in this important characteristic. In the paper under discussion, for example, the 500-watt Coast Guard transmitter employs one modulator tube for one radio power tube, the 2000-watt Coast Guard set employs one modulator tube for two radio tubes, and the 1-kw broadcasting equipment uses four modulator tubes for one radio tube. It is appreciated that these transmitters were produced for different types of applications where the requirements vary greatly and there is no intention to question the conclusions of the designers. The point is that there must be considerable difference in the modulation capabilities of these sets which will be reflected in the range and grade of service obtainable, yet no indication of this discrepancy is conveyed by present rating practices. In view of the growing importance of telephone applications, it would appear that in the interest of technical accuracy this situation should be recognized and corrected without delay.

Under present day conditions there is much to be gained by emphasizing this matter of modulation capability. Considerable evidence is available which indicates that the average modulation capability of the broadcasting stations now in service in this country is not greater than 50 per cent. In other words, the majority of stations cannot modulate more than 50 per cent even during the loudest passages in their programs without serious overloading or "blasting" being observed by their listeners. Having in mind that beat-note interference and power limitation is the most serious problem that confronts the broadcasting industry today, it is of interest to note that the same side band power and, therefore, the same signal-to-noise ratio, could be produced by stations of one-fourth the carrier power output provided their modulation capabilities were doubled, that is, made to approach 100 per cent. Or, to consider another alternative, if the carrier outputs were allowed to remain as they are and the modulation were doubled, the service areas of the majority of stations would be increased by a factor approximating four without increase in beatnote interference. Many listeners would, in fact, experience a noteworthy improvement in beatnote conditions since they would then come within the increased service areas of the neighboring stations. It is not intended to imply that the improvement to be had in this manner is a panacea for our present ills, but since it promises to enable 1-kw stations to afford the same grade of service that we now expect from most 4-kw installations, without increase in the beatnote interference zone, it deserves serious consideration from all factors in the industry.

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* Presented before New York meeting of the Institute, April 4, 1928. Proc. I. R. E., 16, 614; May, 1928.

† Bell Telephone Laboratories, Inc., New York City

The design of a high quality transmitter capable of complete (100 per cent) modulation is a problem of considerable difficulty. With the Heising system, which is now generally employed on account of its lack of critical adjustments and its excellent characteristics with respect to fidelity, the requirement for complete modulation is that the peak value of the alternating voltage superimposed upon the direct plate voltage impressed on the oscillator or modulated amplifier shall be equal to the direct voltage. Under these conditions the instantaneous peak voltage during the positive half of the cycle is twice the direct voltage and during the negative half of the cycle, the instantaneous voltage reaches zero. With the usual circuit arrangement in which the modulator and radio tubes are fed through a common audio-frequency choke coil, complete modulation without serious distortion is obviously impossible, regardless of the number of modulator tubes employed, since it requires that an audio-frequency amplifier (the modulator) supply an alternating voltage equal to the direct voltage impressed on its plate. An obvious solution for this difficulty lies in impressing on the modulator tubes a higher plate voltage than is employed on the radio tubes. This may be accomplished by inserting a suitable resistance in series with the plates of the radio tubes together with a relatively large condenser in shunt to by-pass the audio-frequency component. A further requirement is that ample modulator capacity be provided to satisfy the power conditions. It is well known that the power in a completely modulated wave is 50 per cent greater than that represented by the unmodulated carrier. This power must necessarily be supplied by the modulator together with whatever losses are involved in conversion. With the relatively low efficiencies obtainable in audio-frequency amplifiers, a ratio of modulator capacity to oscillator capacity as high as three or five to one may be necessary. These conditions favor that type of system in which the modulation is effected at low power levels and the requisite power output obtained by subsequent power stages amplifying modulated radio-frequency power.

BOOK REVIEW

Handbook of Chemistry and Physics, a Ready-Reference Pocket Book of Chemical and Physical Data. By Charles D. Hodgman and Norbert A. Lange. Twelfth edition. Chemical Rubber Publishing Company, Cleveland, Ohio. 1112 pages. Price $5.00.

This handbook of chemical and physical data is too well established as a valuable aid in the laboratory and designing room to require a lengthy review. Although none of the new material is of special interest to the radio engineer, still he will find the volume as a whole to be a true "handbook," and not a mere shelf-book. Among the subdivisions that are likely to be of most service from the radio point of view are the mathematical tables, wire tables, units and conversion factors, definitions, laws and formulas of physics, properties of conductors and dielectrics, laboratory arts and recipes and numerical data on densities, elasticities, strength of materials, thermal expansion, etc. About ten pages are devoted to radio formulas taken from the Bureau of Standards publication on radio instruments and measurements.

W. G. Cady †

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Volume 16, Number 12 *December, 1928*

MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE*

THIS is a monthly list of references prepared by the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The articles listed below are not obtainable from the Government. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R100. Radio Principles

R112.1 Gratsiatos, J. Über das Verhalten der radiotelegraphischen Wellen in der Umgebung des Gegenpunktes der Antenne und über die Analogie zu den Poissonschen Beugungserscheinungen. (On the behavior of radio waves in the vicinity of the image point of the antenna and on the analogy to refraction phenomena due to Poisson.) *Annalen der Physik*, 86, 1041–1061; August, 1928.

(The electric potential at the image point of a transmitting set and in the vicinity is derived based on the theory of Watson. This gives expressions for the electric and magnetic field.)

R124 Nestel, W. Untersuchung der Brauchbarkeit von Rahmenantennen fur Sendezwecke. (Investigation of the usefulness of coil antennas as transmitters). *Zeitschrift für Techn. Physik*, 9, 143–145; 1928.

The radiation efficiency of coil antennas is investigated by means of Rudenberg's formulas. It is shown that for short waves the coil antenna is almost as efficient as ordinary antennas.)

R125.6 Wilmotte, R. M. General considerations of the directivity of beam systems. *Journal* Institution of Elec. Engrs. (London), 66, 955–961; September, 1928.

(Definitions are given for directive efficiency and sharpness of directivity in order to treat theoretically the best condition for an effective beam system. An inclined antenna system with a reflector is suggested and an improvement on the Franklin antenna.)

R125.6 Wilmotte, R. M. and McPetrie, J. S. A theoretical investigation of the phase relations in beam systems. *Journal* Institution of Elec. Engrs. (London), 66, 949–54; September, 1928.

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* Original Manuscript Received by the Institute, October 15, 1928.

(The authors derive expressions for the phase relation of beam systems They assume that the field at all points is due to the radiation field and apply to the amplitude and phase certain factors taking the distance into account. The factors can be read off a graph and their results are checked against experimental investigations due to Tatarinoff.)

R127 Wilmotte, R. M. The nature of the field in the neighborhood of an antenna. *Journal* Institution of Elec. Engrs. (London), 66, 961–67; September, 1928.

(Methods for the calculation of the induced voltage in a receiving antenna in the neighborhood of a transmitting station are given.)

R131 Scroggie, M. G. A direct-reading valve tester. *Experimental Wireless* (London), 5, 480–84; September, 1928.

(An apparatus is described for the direct indication of the mutual conductance and anode resistance of electron tubes.)

R132 Beatty, R. T. The stability of a valve amplifier with tuned circuits and internal reaction. Physical Society *Proc.* (London), 40, 261–268; August 15, 1928.

(Algebraic and graphical treatment of tuned circuit amplifiers.)

R133 Wechsung, W. Die Erzeugung sehr kurzer elektrischer Wellen mit Wechselspannung nach der Methode von Barkhausen und Kurz. (The production of very short electric waves with alternating current by the method of Barkhausen and Kurz.) *Zeitschrift für Hochfrequenztechnik,* 32, 58–65; August, 1928.

(A continuation of a paper appearing in the *Jahrbuch*, p. 15, 1928. The author extends his experimental work to Barkhausen oscillations with alternating current excitations).

R134 A new idea for a detector valve. *Experimental Wireless* (London), 5, 515; September, 1928.

(The separation of slower moving electrons from the faster ones by means of a magnetic field and a special grid is suggested. This would give rise to a more sensitive detector tube.)

R140 van der Pol, B. and Van der Mark, J. Le battement du coeur considere comme oscillation de relaxation et un modele electrique du coeur. (The beating of the heart considered as relaxation oscillation and an electric model of the heart). *L'Onde Electrique,* 7, 365–392; September, 1928.

(A review of work on relaxation oscillations and a brief description of the system of frequency division. Compares the action of the heart beat with that of a neon tube oscillator showing that the period of the heart beat has more or less a relaxation period.)

R150 Sixtus, K. Über den Schwingkristall. (On the oscillating crystal). *Zeitschrift für Techn. Physik,* 9, 70–74; 1928.

(From the static voltage current characteristic of a contact detector it is concluded that oscillations are produced when working in the falling portion of the characteristic. A theoretical formula based only on heating effects at the point of contact confirms the experiment.)

R190 Eccles, W. H. and Leyshon, W. A. Some new methods of linking mechanical and electrical vibrations. Physical Society *Proc.* (London), 40, 229–233; August 15, 1928.

(Circuits are shown for which contact detector and neon tube oscillations are controlled either by a tuning fork or a quartz plate.)

R200. RADIO MEASUREMENTS AND STANDARDIZATION

R214 Hitchcock, R. C. Piezo-electric frequency control. *Electric Journal,* 25, p. 503; October, 1928.

(An account of the frequency of the station KDKA during the past half year.)

R283 Symonds, A..A. Loop permeability in iron, and the optimum air gap in an iron choke with d.c. excitation. *Experimental Wireless* (London), 5, 485–490; September, 1928.

(Experimental study of the incremental permeability of iron when d.c. magnetisation is superimposed.)

R284.3 Meissner, A. and Bechman, R. Untersuchung und Theorie der Pyroelektrizität. (Investigation and theory of pyro-electricity). *Zeitschrift für Techn. Physik*, 9, 175–85; 1928.

(Experimental and theoretical investigation of the pyroelectric effects of quartz and Turmalin.)

R300. RADIO APPARATUS AND EQUIPMENT

R342 von Ardenne, M. and Stoff, W. The harmful effects of interelectrode capacity. *Experimental Wireless* (London), 5, 509–514; September, 1928.

(Description of the well-known effects of the tube capacity in amplifiers.)

R342.2 Barclay, W. A. A graphical construction for resistance amplifiers. *Experimental Wireless* (London), 5, 499–500; September, 1928.

(A graphical method for determining the best values of anode resistance, grid bias, etc., of an amplifier.)

R342.2 Ramelet, E. Über die neue rein elektronische Verstärkung verwendende Zählmethode für Korpuskularstrahlen. (On a new purely electronic amplification for counting corpuscles). *Annalen der Physik*, 86, 871–913; August, 1928.

(The pure electronic amplification (not amplification by impact ionization as originally employed by Rutherford-Geiger) due to Greinacher has been worked out in more detail by means of a suitable resistance-capacity coupled amplifier.)

R344 Hollman, H. E. Ein Röhrenoszillator für sehr kurze ungedämpfte Wellen. (A tube generator for very short waves.) *Annalen der Physik*, 86, 1062–1070; August, 1928.

(A tube generating set is described which works down to 36 cm. In some cases waves down to 13.2 cm were measured.)

R374 Ogawa, W., Nemoto, C., and Kaneko, S. The effect of chemical composition on the sensitivity of galena as a radiodetector and the cold emission from crystals. Researches of the Electrotechnical Laboratory, Japan, No. 230, June, 1928.

(The author explains the action of the crystal detector by means of the difference of electron emissions from the two electrodes forming the contact.)

R376 Doucet, V. La distortion dans les ecouteurs telephoniques. (On the distortion in telephone receivers). *QST Français et Radioelectricite Reunis*, 9, 27–29; September, 1928.

(Analytical investigation of telephone receivers with regard to distortion.)

R376 Bauder, R. and Ebinger, A. Untersuchungen über Monotelephone. (Investigations on the Monotelephone). *Zeitschrift fur Techn. Physik*, 9, 65–69; 1928.

(Based on the theory of a telephone receiver whose stationary field is large in comparison to the a.c. field. Experiments are carried on with mechanical tuning to resonance. Interesting photographs are shown of the nodal lines developed on the diaphragm.)

R376.3 Toulon, P. L'evolution et l'avenir de haut-parleurs, exemples de principes nouveaux: Haut-parleurs electro-statiques. (Evolution and future of loudspeakers, examples of new principles: electrostatic loudspeakers). *L'Onde Electrique*, **7**, 393–409; September, 1928.

(Remarks on present day loudspeakers and a description of an electrostatic type.)

R376.3 Clark, H. A. and Bligh, N. R. Some output power measurements on a moving coil drive loudspeaker. *Experimental Wireless* (London), **5**, 491–98; September, 1928.

(Experimental method (using the Heaviside equal ratio inductance bridge) for determining the resistance and reactance of a loudspeaker in the audible range.)

R384.1 Hull, R. A. The frequency measurement problem. *QST*, **12**, 9–19; October, 1928.

(Description of frequency meter and monitor for 1929 to be used by amateurs in keeping their transmitting stations on their proper frequency.)

R386 Winter-Gunther, H. Zur Theorie der Siebketten. (On the theory of filters). *Zeitschrift für Hochfrequenztechnik*, **32**, 41–46; August, 1928.

(A treatment of filter circuits from the standpoint of coupled circuits. Instead of the method due to H. Riegger and L. Cohen, the method of normal coordinates as used by Routh and Lord Rayleigh for mechanical systems is introduced.)

R386 Mallet, E. Chains of resonant circuits. *Journal* Institution of Elec. Engrs. (London), **66**, 968–74; September, 1928.

(The coupled circuit filter is solved with the method of differential equations in order to obtain an expression for the current in the last link of the chain. A graphical solution with an example is given.)

R500. APPLICATIONS OF RADIO

R580 Taylor, J., and Taylor, W. Some new applications of short radio waves. *Experimental Wireless* (London), **5**, 503–508; September, 1928.

(Shows experiments in discharge tubes when high-frequency voltages are applied. It is possible to produce discharges even at very low pressures.)

R800. NON-RADIO SUBJECTS

517 Berg, E. J. Heaviside's operational calculus as applied to engineering and physics. *General Electric Review*, **31**, 504–509; September, 1928.

(The concluding section of a series of papers on above subject. It gives an application to the linear flow of heat and compiles the formulas used in operational calculus.)

534 Lindsay, R. B. High-frequency sound radiation from a diaphragm. *Physical Review*, **32**, 515–519; September, 1928.

(From the standpoint of hydrodynamics and sound, an expression is derived for the intensity of the high-frequency sound radiating from piston-like oscillator at a distance from the oscillator greater than the diameter of the circular radiator.)

534 Brenzinger, M., and Dessauer, F. Eine neue Methode unmittelbarer Steuerung der Luft durch elektrische Schwingungen. (A new method for the direct control of air waves by means of electric oscillations). *Physikalische Zeitschrift*, **29**, 654–58; September 15, 1928.

(A glow discharge is used for changing directly superimposed alternating currents into sound waves. The same principle is also used for making a glow discharge microphone.)

534 Meyer, E., and Just, P. Zur Messung von Nachhalldauer und Schallabsorption. (On the measurement of the reverberation time and sound absorption). *Elektrische-Nachrichten Technik*, 5, 293–300; August, 1928.

(A method is given for determining the reverberation time. The experimental curves prove the exponential decay of the sound for an echo.)

535.3 Koller, L. R., and Breeding, H. A. Characteristics of photoelectric tubes. *General Electric Review*, 31, 476–79; September, 1928.

(Characteristic curves of light, current, sensitivity, and gas pressure for photoelectric cells.)

537.66 Neidl, G. Neuer Versuch zum Johnson-Rahbeck-Effekt. (New experiment with the Johnson-Rahbeck effect). *Zeitschrift für Techn. Physik*, 9, 22; 1928.

(A sphere of mercury is placed on a half conductor, the lower face of which has a metal coating. When an alternating voltage is applied to the mercury and the metal coating, the mercury will vibrate in synchronism with the alternating voltage.)

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED
November 7, 1928

Transferred to the Fellow grade

New York	New York City, 282 West End Avenue	Dreher, Carl

Transferred to the Member grade

California	Hollywood, KMTR Radio Corp., 1025 Highland Avenue	Van Why, Forbes William
Illinois	Chicago, University of Chicago, Ryerson Lab.	Hoag, J. Barton
New Jersey	Madison, 14 Fairwoods Road	Manning, Charles T.
England	London, W. 12, Emlyn Road	Hardy, Harold

Elected to the Member grade

Dist. of Columbia	Washington, 4302 Brandywine St. N.W.	Davis, Thomas McL.
Massachusetts	W. Somerville, 53 Francesca Avenue	Kolster, Charles C.

Elected to the Associate grade

Alabama	Phenix City	Moore, C. M.
California	Berkeley, 2112 Addison	Howard, L. W.
	Burlingame, 908 Morrell Avenue	Michalic, Charles Tom
	Hollywood, 1301 Martel Avenue	Kratz, C. F.
	Marshall, c/o Radio Corporation of America	Lawrence, Lloyd H.
	Oakland, 643 Poirier St.	O'Brien, Daniel L., Jr.
	Orange, P. O. Box 145	Charbonneau, L. H.
	San Francisco, 140 New Montgomery, Room 1024	Stewart, Ronald B.
	San Francisco, KFWI, 1182 Market St., Room 203	Weir, F. Torres
	San Pedro, U. S. S. *Procyon*	Granum, Alfred M.
	Santa Ana, 1334 S. Parton St.	Sleeper, J. Lloyd
Georgia	Atlanta, 684 Durant Place	Gue, E. M.
	Atlanta, c/o Radio Station WSB	Shropshire, A. W.
	Columbus, 731 Broad St.	Cowart, Frank P.
	Columbus, South Georgia Power Co.	Davis, James C.
Illinois	Chicago, 3814 Madison St.	Haire, A. F.
	Chicago, 1119 So. Wood St.	Korbel, George William
	Chicago, c/o R. C. A., 100 W. Monroe St.	McCune, R. H.
	Chicago, 4137 Jackson Blvd.	Strandjord, Millard F.
	Chicago, 29 No. Morgan Street	Trier, Stanley S.
	Rantoul, Chanute Field, Communications Section	Fiechtner, Theodore G.
	Rockford, 1506 School St.	Johnson, Art A.
Kansas	Atchison, 202 N. 2nd St.	Dimond, Benjamin Dan
Louisiana	New Orleans, 528 N. Hennessy St.	Salzer, Herman M.
Maine	Belfast, c/o R. C. A.	Morris, Hedley B.
Massachusetts	Braintree, 149 Hollis Ave.	Robison, Philip F.
	Cambridge, Harvard University, Pierce Hall, Room 204B	Samoiloff, Leon
	Chatham, Box 631	Robinson, Forrest D.
	Chathamport, Radio Hotel	Lee, Walter O.
	North Cambridge, 128 Clay Street	Grenier, Henry N.
	Medford, 16 Higgins Avenue	Yorks, William Ray
	South Boston, 157 N. Street	Brown, George W.
Michigan	Crystal Falls, 225 Superior Ave.	Lee, Raymond R.
Missouri	Kansas City, 2803 Linwood Blvd.	Schulze, Heri T.
Nebraska	Norfolk, P. O. Box 226	Leeman, Wilson
New Jersey	Cliffwood, Box 31	Jansky, Karl G.
	East Orange, 106 North Walnut St.	Edison, Theodore M.
	Long Branch, 58 Washington St.	Watson, Paul Edwin
	Weehawken, 2 Fourth St.	Girard, E. J.

New York	Brooklyn, 896 Troy Avenue	Riccobono, Sebastian
	Buffalo, 20 Ullman St.	Mahler, George P.
	East Rockaway, L. I., 47 Waldo Ave.	Van Duyne, Eugene Douglas
	Kenmore, 37 Princeton Blvd.	Voll, Harry F.
	New York City, 161 West 140th Street	Cox, Lloyd V.
	New York City, 562 W. 183rd Street	Gati, Bela
	New York City, 50 W. 57th Street	Leonard, John M.
	New York City, Mackay Radio and Tel. Co., 67 Broad Street	Nivison, T. E.
	New York City, 19 Grove Street	Reardon, Daniel
	Riverhead, L. I., P. O. Box 1077	Tammaro, Joseph Carl
	Rocky Point, Radio Central	Draigh, Canton V.
	Schenectady, 172 Nott Terrace	Sohon, Harry
	Schenectady, 9 N. Church Street	Woodworth, John L.
North Carolina	Charlotte, Y.M.C.A.	Saylor, J. G.
Ohio	Cleveland, 1240 East 167th Street	Shaw, John Joseph
	Fostoria, 642 Lynn Street	Elsea, Farrell F.
	Shiloh, Box 112	White, Kenneth E.
Pennsylvania	Allentown, 806 Union Street	Thomas, Chas. V.
	Pittsburgh, West Penn. Bldg., 14 Wood Street, Room 1304	Sunnergren, Arvid P.
	Waverly, Box 163	Mach, Dahl W.
	Wilkes-Barre, 429 George Ave.	Speicher, Ellsworth J.
Rhode Island	Providence, 42 Greenwich Street	Maker, Albert Edwin
South Dakota	Yankton, 318 W. First Street	Seils, Harry A.
Tennessee	Knoxville, P. O. Box 531	Adcock, S. E.
	Memphis, 43 So. Barksdale	Dowler, R. B.
Texas	Dallas, 3106 St. John Drive	Hinsch, Lincoln A.
	Livingston	McClanahan, C.
Washington	Seattle, 7708 Latona Ave.	Brandt, Oscar T. D.
	Seattle, 904 Telephone Bldg.	Budden, F. W.
	Seattle, 3234 Belvidere Ave.	Hamilton, Edward A.
	Seattle, Pacific Tel. and Tel. Co.	Schreiber, Ernst H.
	Seattle, 4026 Evanston Ave.	Sletmoe, A. M.
West Virginia	McMechen, 1104 Caldwell St.	Hicks, William Ray
Wisconsin	Watertown, 314 Water St.	Ebert, Sylvanus J.
Australia	Melbourne, East Camberwell, 6 Beech St.	Fitts, Rupert Alfred
Canada	Victoria, B. C., 2084 Newton St.	Hawkins, Ernest
	Ottawa, 21 Florence St.	Donaldson, Bruce W.
China	Hong Kong, c/o Electrical Dept. P.W.D.	Logan, James Stanley
England	Derbyshire, Buxton, 2 Wood Cliffe	Smith, Harold Ingleby
	Middlesex, Shipperton, Pharaoh's Island, "Kantara"	Henderson, Walter B.
Japan	Shizuoka-Ken, Ogasagun, Kakegawa-cho Kandaiji	Yokoyama, Tetsumi

Elected to the Junior grade

California	Los Angeles, 1311 Citrus, Hollywood	Fox, B. M.
Nebraska	Clay Center	Hertel, Roger H.
New York	Brooklyn, 6 Bay 23rd St.	Liebner, Barney
	Buffalo, 78 East St.	Ferger, Herbert
	Jamestown, 200 Hallock St.	Ellis, James G., Jr.
	Lancaster, 69 Aurora St.	Pictor, Robert Ezra
Pennsylvania	Altoona, 602 E. Grant Ave.	Youngkin, E. E.
	Philadelphia, c/o Atwater Kent Co.	Reid, Floyd Freeman
	Philadelphia, 5737 N. Lawrence St.	Trumpy, J. Walter
	Scranton, 1618 Monsey Ave.	Bartzel, Charles R.
Wisconsin	Milwaukee, 1278 W. 24th St.	Moorbeck, Clinton
England	Manchester, Trafford Park, Metro-Vickers	Thomas, Guy Henry
	Wiltshire, Swindon, 7, Pinehurst Road	Humphreys, Leonard W.

APPLICATIONS FOR MEMBERSHIP

Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Committee on Admissions. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before December 29, 1928. These applications will be considered by the Board of Direction at its January 2, 1929 meeting.

For Election to the Fellow grade

Dist. of Columbia	Washington, Naval Communications, Navy Dept	Craven, T. A. M.

For Transfer to the Member grade

Michigan	Muskegon, 1390 Palmer Avenue	Richardson, Avery G.
New York	New York City, 195 Broadway Room 1607	Misenheimer, Harvey N.
Hawaii	Hilo, Box 923	Branch, L. W.

For Election to the Member grade

Massachusetts	Cambridge, 50 Kirkland Street	Black, K. Charlton
New Jersey	Bloomfield, 120 Berkeley Avenue	Morehouse, W. B.
Germany	Berlin-Charlottenburg, Cauerstrasse 19III	Kofes, Albert

For Election to the Associate grade

Arkansas	Camden, Box 661	Tagart, Sam W.
	Texarkana, 2119 Hickory Street	Wallace, L. E.
California	Los Angeles, 947 Francisco Street	Heimberger, Albert E.
	San Francisco, 378 Golden Gate Ave., Apt. 235	Kelsey, Karl D.
Connecticut	Hartford, 33 Eastview Street	Le Conche, Carl
Dist. of Columbia	Washington, 3131 Newton Street, N.E.	Carroll, Thomas D.
Illinois	Chicago, 4141 Fifth Avenue	Felio, Orin J.
	Chicago, 2530 Kedzie Blvd.	Krueger, Alfred
	Chicago, 2530 Kedzie Blvd.	Wilm, C. F.
	Galesburgh, 895 West Main Street	Mead, Leo R.
Indiana	North Manchester, 702 North Walnut St.	Grove, Claude C.
	Richmond, Box 6, Earlham College	Hickman, Roger W.
	Valparaiso, 825 Lincolnway	Swanson, Carl R.
Kansas	Topeka, c/o Radio Station WIBW	Herider, Ernest D.
	Wichita, 104 Severdale Apts.	Mitchell, Ray H.
Kentucky	Covington, 3410 Church Street	Fraasa, C. F.
Louisiana	New Orleans, c/o Tropical Radio Tel. Co. 321 St. Charles Street	Allston, W. F.
	Port Barre	Futral, John E.
Maine	Bangor, 331 Center Street	Creamer, W. J., Jr.
Massachusetts	Boston, 200 Huntington Avenue	Browne, Monte C.
	Boston, 468 Massachusetts Ave.	Wass, Howard H.
	Cambridge, 28 Gorham Street	Shen, Pao-guay
	Cambridge, 28 Gorham Street	Tsao, T. C.
	Rockport, 7 Gott Street	Mills, William P.
	Springfield, 38 Greenacre Square	Nystrom, Raymond A.
Michigan	Ann Arbor, 520 Forest Street	Nelson, C. Emory
	Jackson, 1019 First Street	Planck, R. M.
Mississippi	Mississippi City, P. O. Box 117	Pecoul, Ferdinand Anthony
Montana	Forsyth, Box 1064	Roberson, Carl
Nevada	Reno, University of Nevada	Sandorf, Irving Jesse
New Jersey	Bridgeton, 150 East Avenue	Nichols, Howard Leslie
	East Orange, 71 Leslie Street	Woodworth, Fred B.
	Jersey City, 169 Zabriskie Street	Weber, Walter
	Orange, 454 Conover Terrace	Kynor, Merrill W.
	Union City, 301–33rd Street	Masalkovics, Joseph Albert
New York	Brooklyn, 186 Hopkinson Avenue	Potter, Max
	Brooklyn, 164 Columbia Heights	Stinchfield, J. Maxwell
	Buffalo, 46 Riley Street	Oversmith, La Rue H.
	Buffalo, 605 Auburn Avenue	Perau, Frederic Henry

New York (cont'd.)	Buffalo, 62 Manden Street	Smith, Stanley C.
	Gardenville, 447 Potter Road	Felmet, Albert
	New York City, c/o Norton Lilly, 26 Beaver Street	Berry, Harold C.
	New York City, 140 West 16th Street	Castaneda, Santiago
	New York City, 118 East 103rd Street	Doubles, David J.
	New York City, 41 West 86th Street	Feldstein, Martin A.
	New York City, c/o Norton Lilly, 26 Beaver	Francis, Philip
	Brooklyn, 326 Utica Avenue	Gottdenker, Martin
	New York City, R. C. A., 326 Broadway	O'Connor, John G.
	New York City, 195 Broadway	Reinken, Louis W.
	Rochester, 7 Edmonds Street	Steneri, Arthur John
	Schenectady, 848 Union Street	O'Neill, John P.
	Staten Island, USS *Acushnet*	Miller, Paul E.
North Carolina	Henderson	Woolard, E. W.
North Dakota	Fargo, 905-5th St., North, No. 19 Rust Apts.	Cook, Tedd W.
Ohio	Cleveland, 7617 Myron Avenue	Crocus, John Stanley
	Cleveland, 3516 Storer Avenue	Irvine, Robert P.
	Marion, 177 East Center Street	Brown, David Ashton
	Zanesville, 606 Ridge Avenue	Bell, Lewis M.
Oklahoma	Oklahoma City, Oklahoma Gas & Electric Co.	Bathe, C. E.
Oregon	Portland, 873 Belmont Street	McCargar, S. Harold
Pennsylvania	Allentown, 246 S. Madison Street	Bowman, Charles W.
	Allentown, 226 N. 6th Street	Haines, A. J. D.
	Allentown, 2014 Highland Avenue	Muthart, John A.
	Allentown, 949 Hamilton Street	Rauhofer, Frank
	Easton, 426 S. 21st Street	Clendaniel, John Edwin
	Easton, 18 N. 9th Street	Raesly, James B.
	Easton, 1815 Fairview Avenue	Weller, Everett Clare
	Philadelphia, 2209 South Chadwick Street	Johnson, Harmon
	Phialdelphia, 765 S. 10th Street	Mattia, Ralph F.
	Wilkinsburg, 215 Ross Avenue	Reynolds, C. C.
Rhode Island	Providence, 4 Pemberton Street	Brewster, O. H.
West Virgina	Charleston, 4 Maple Terrace	Moore, Thomas H.
Wisconsin	Milwaukee, 1667 Oakland Avenue	Hough, W. E.
	Milwaukee, 1405 Bremen Street	Strassman, Irving H.
British Guiana	Demerara, Georgetown, 61 Hadfield St.	Tasker, Joseph Thomas
Chile	Valparaiso, Casilla 1653	Vierling, Gustav
England	Stockport, North Reddish, 446 Gorton Road	Howard, C. Alexander
India	Baroda, Babajipura	Dighe, K. S.
Ireland	60 Clifton Road, Bangor, Co. Down	Jamison, A.
Mexico	Puerto Mexico, Veracruz, Apartado, 86	Bourgeois, Allen B.

For Election to the Junior grade

Illinois	Champaign, 61 East Green Street	Pennington, D. J.
Indiana	Valparaiso, 712 Calumet Avenue	Clark, Edgar J.
	Valparaiso, 156 S. Franklin Street	Woodworth, Elwyn Crane
Massachusetts	Quincy, 189 Common Street	D'Alessandro, Genaro
Michigan	Royal Oak, 286 West Ten Mile Road	Alain, J. E.
New York	Brooklyn, 140 Vanderbilt Avenue	McGonigle, William J.
Oregon	Salem, 2405 Center Street	Poujade, Donald G.
Pennsylvania	Allentown, 38 S. Jefferson Street	Foley, W. R.

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